REPORT OF ... INDUSTRIAL EXPERIENCE TOUR

G. W. CLARKE

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REPORT OF ... INDUSTRIAL EXPERIENCE TOUR

G. W. CLARKE

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Thesis
C492

LCDR G. W. Clarke, USN
Westinghouse Electric Corporation
2519 Wilkens Avenue
Baltimore, Maryland
March 21, 1952

=Report of an industrial experience tour=

Dr. Austin R. Frey
U.S. Naval Post Graduate School
Monterey, California

Dear Sir:

Enclosed is the report covering my activities during the thirteen week industrial experience tour as required for completion of the curriculum in Engineering Electronics.

My opinion of this final phase of the program of postgraduate work is very high. It among other things clearly points out the respect due those who developed the techniques and engineering concepts so ably taught us while in the school. I consider it an essential phase for a fully rounded engineering education.

Sincerely,

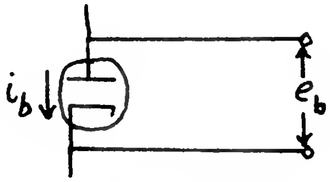
IONIZATION CHAMBER DIRECT COUPLED LOGARITHMIC AMPLIFIER

The first problem proposed was a study of direct coupled (hereinafter DC) amplifiers as applied to measurement of currents thru the range of 10^{-12} thru 10^{-4} amperes. The scope of the problem was somewhat limited since the input device (source) was to be considered as a constant current one, i.e.: one of infinite internal impedance. The drift stability aim of this study was a DC amplifier having a maximum of $\pm$ 5% drift where drift is defined as incremental current change divided by the absolute value of the current being read at the time. Since the drift of the amplifier had to be measured with reference to an absolute value of current input and the measurement of the absolute value of the input depended upon the characteristics of the input device which was in this case the log diode (discussed below), the problem became a study of a DC amplifier with a log-diode input. This is discussed in detail in App I.

The original work lay in a search of the literature in regards to the limitations of direct coupled amplifiers and existing solutions to the problems involved. The results of this search are affixed in Appendix III. Further study of the literature in regards to logarithmic characteristic current measurement devices was considered unnecessary since it is well known that of all direct current logarithmic devices the only one sufficiently temperature insensitive with a great enough range (8-9 decades) is the high vacuum diode operated at reduced filament voltage (hereinafter - log diode).

In addition to the limitations of the basic DC amplifier as discussed in App. III there is imposed in measurement of currents of the order of 10^{-12} ampere the consideration of grid current. (The log diode is essentially a high impedance device, its differential impedance being derived as $R = \frac{0.1}{I}$ where R is in ohms and I is the current thru the diode in amperes.

From



$$e_b = K \log_{10} i_b + \text{constant.}$$

For ordinary small diodes experiment yields

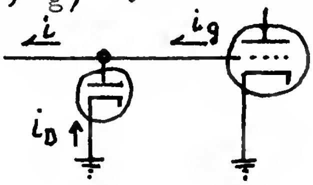
$$K = 0.25 \text{ volt/decade of current.}$$

$$r_p = \frac{\partial e_b}{\partial i_b} = \frac{d\, e_b}{d\, i_b} =$$

$$= \frac{d\,(K \log_{10} i_b + \text{constant})}{di_b} = \frac{K \log_{10} e}{i_b}$$

$$r_p = \frac{0.434 \times 0.25}{i_b} \approx \frac{0.1}{I_b} \qquad)$$

If the grid current thru the diode load tube be of the order of 10^{-14} ampere then grid current, i_g, may be considered negligible



as compared to diode current, i_D, for measurement purposes. This consideration

limited selection of DC amplifier input tubes to electrometer tubes. (The use of a vacuum tube in electrometer applications is proposed by Metcalf & Thompson.

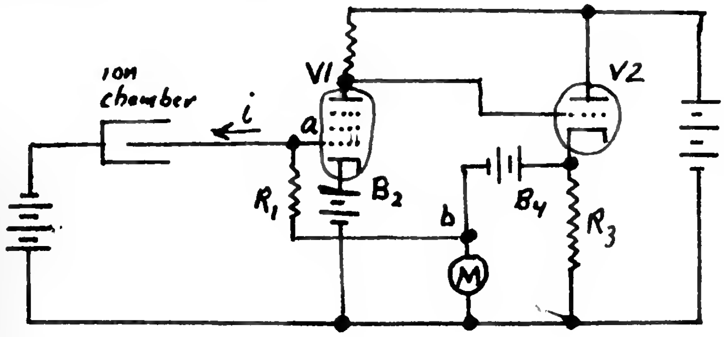
Consideration is taken grid current sources;

1. Leakage over glass and insulating supports.
2. Positive ions due to ionization of residual gas.
3. Thermionic emission from grid due to heating of grid by cathode power.
4. Positive ions emitted by cathode.
5. Photoelectrons emitted by control grid due to light from cathode.
6. Photoelectrons emitted by control grid due to soft X-rays produced by plate current electrons striking plate.

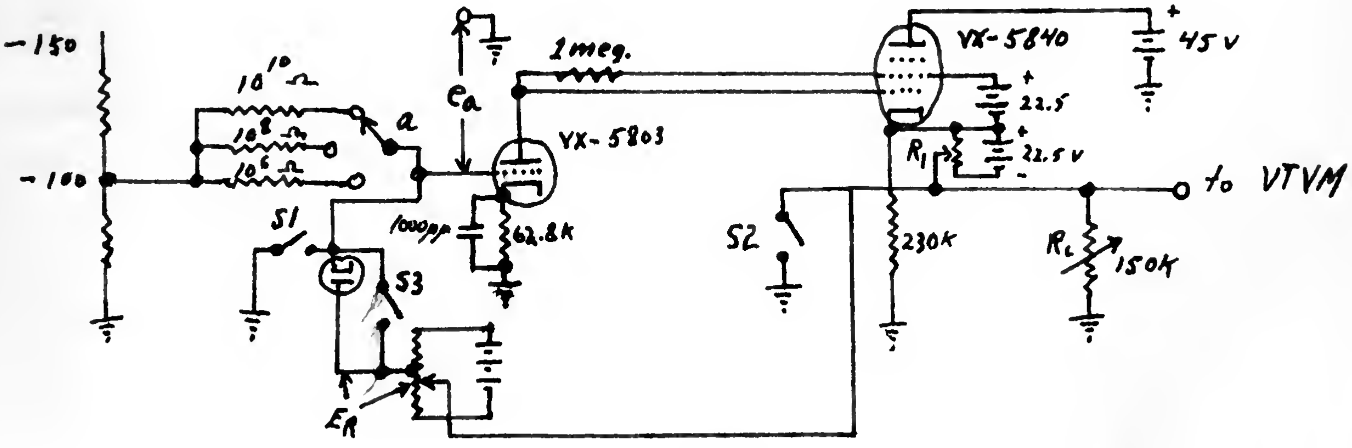
Corrective constructional and operating steps are described. The below two commercially available tube types are constructed and operated in accordance with the above considerations. Typical types of electrometer tubes are the Raytheon CK 571 AX pentode rated at 10^{-14} ampere grid current for 0.25 volt input or the Victoreen VX-5803 triode rated at 10^{-14} ampere at -0.5 volt input. Unavailability of the Raytheon type limited experimentation to Victoreen VX-5803 triodes.

The first laboratory set-up originated in study of a circuit devised to minimize effects of contact potential (work function) variations in the electrometer tube in creating drift. (Contact potential is the term applied to the potential between two pieces of metal not in contact due to differences of work function). The reasoning employed showed that if the input voltage to the

electrometer tube could be greatly increased without exceeding the allowable input voltage, from grid current considerations, then the contact potential effect would be relatively negligible.



Assuming the cathode follower efficiency 100% then the gain with feedback of 100% is $\frac{1}{1-A}$ or the input resistor R1 may be increased in value by a factor of 1-A where A is V1 gain. This change with a given current i gives the desired result of an effective increase of input voltage while remaining within grid current limitations. Note: no attempt is made to provide for drift due to variations of the value of R1. The construction was made with a log diode (Sylvania 5647) in place of R2 to provide several decades coverage of input current without resort to switching and without exceeding the electrometer tube limits. The tubes used were those readily available.



Since efficiency of the cathode follower $\eta = \dfrac{R_c}{\frac{1}{g_m} + R_c}$

or $\eta = \dfrac{G}{1 + G}$ where $G = g_m R_c$, R_c is made variable. The reasoning followed was that if

$$\eta A = 1$$

then the input point "a" could always be returned to ground potential for measurement purposes by means of varying R. The procedure to be followed was

1. Close S1 and S2 and vary R1 until the VTNM indicated zero output voltage.

2. Open S2 close S3 and insert known potential at "a" by means of R.

3. Vary R_c to return e_a to zero. This would provide unity feedback.

4. With S1, S2 and S3 open insert current in loop by selector switch. Read VTVM.

5. Vary R until VTVM reads zero.

6. Read E_R

Since "a" is again at ground potential the current thru the diode is that inserted. The VTVM reading is the output voltage corresponding.

The Victoreen VX-5803 has an amplification factor of 2 and the diode efficiensy was computed as 98%. In spite of considerable manipulation of circuit parameters, however, the gain "A" of the VX-5803 could not be raised sufficiently to obtain a loop gain of 1. Therefore this project was rejected since the entire

measurement technique was based upon ability to return "a" to ground potential.

Subsequent investigation indicated that no diode existed with the differential impedance desirable for use in this circuit in the prescribed current range.

A second attack on the contact potential problem was based on the writers assumption that in a common cathode duo-diode variations in work function due to cathode temperature variations would give rise to equal contact potential variations with both diodes. Such a property, if possessed by a common cathode duo-diode, would be very useful in providing a signal independent of heater potential variations. A check of existing tube types showed no such tube available. The alternatives of using the diode sections of a duo-diode triode or a duo-triode (diode connected) were selected. The details are pointed out in App. I.

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1. Phys. Rev. 36, 1489, 1930.
2. Theory of Thermionic Vacuum Tubes - E.L. Chaffee McGraw-Hill Book Co. 1933.
3. An Improved D.C. Amplifier for Portable Ionization Chamber Instruments, N. F. Moody, Rev. Sci. Inst. April 1951.

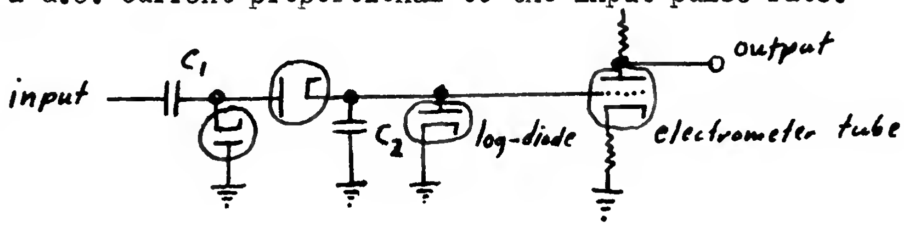
A PULSE DISCRIMINATOR CIRCUIT

The second problem arose out of the design specifications for the ASMC-1 reactor control system which called for a minimum number of vacuum tubes consistent with fission counting accuracy for control purposes.

In a highly accurate system using fission counters the ionization pulse from the counter is amplified and applied to a Schmidt trigger circuit which is followed by a cathode-coupled multivibrator and diode counting circuit. The output of the diode counter is applied to a log-linear diode (described previously) and electrometer tube indicator to indicate pulse rate.

A brief explanation for this choice of circuitry is offered. The fission counter (an argon filled cylinder lined with U-235 concentric with a center electrode across which a high field is impressed) lies in a concentration of alpha particles, gamma rays and thermal neutrons. All of these produce ionization of the argon, the alpha particles and gamma rays directly, the thermal neutrons by fission of the U-235 into heavy ionized particles. The direct ionization current level thm the fission counter due to gamma radiation alone is large compared to that due to alpha particle and fission ionization. Hence, at the reactor power levels where fission counting techniques are used, pulse amplification is used to obtain information of pile period (rate of change of thermal neutron level) and level rather than direct ionization current. Discrimation by means of pulse height is used to separate

pulses due to fission ionization from those due to alpha particle ionization. (Note: These fission counters are operated with gas densities and electric fields so adjusted as to give a "gas amplification" of one i.e.: a single positive ion produced at the plate will arrive at the cathode without causing collision ionization due to the fact that its acceleration in the field is insufficient to bring it to collision ionization velocity in the distance traversed. The fission particles from the U-235 have velocities ranging from that insufficient to cause collision ionization to much greater velocities. Hence, most of these particles produce larger output pulses than the alpha particles which have, on the average less velocities.) The distribution of these fission pulses is on the average ten microseconds between pulses although satistical variations indicate pulse separations of less than one microsecond. A choice of pulse resolution time of one microsecond for the amplification and discrimation circuitry was chosen as it leads to less than 5% error in pulse rate indication. The output of the amplifier discriminator feeds a diode counting circuit whose output is a d.c. current proportional to the input pulse rate.



C_2 is very much greater than C_1 and therefore the designation "bucket" is applied to C_1 since it governs the charge transferred to C_2 per pulse.

The potential across the log-diode is the logarithm of the current thru it. The current thru the diode is governed by the potential across C_2. This log-linear measurement is necessary since the pile thermal neutron level varies exponentially. Note that input pulse width in this circuitry (within large limits) has no effect on the charge per pulse transferred to C_2 and hence counting rate. However, the input pulse amplitude has a decided effect. Therefore our amplifier-discriminator circuitry must provide pulses of constant output amplitude for all fission counter pulses above a chosen pulse height. A further consideration in the choice of discriminator is that the tripping level must be capable of accurat setting.

The requirements on the system are summed up therefore as:

1. Accurately adjustable pulse height discrimination.
2. One microsecond resolution of pulses.
3. Output pulse amplitude independent of input pulse amplitude.

The Schmidt trigger circuit is capable of satisfying the first two of the above requirements and, when followed by a cathode-coupled multivibrator, all three. But, the consideration of reduction in number of vacuum tubes led to use only of the Schmidt circuit.

The problem facing the writer was to investigate the behavior of the Schmidt circuit as a function of its parameters. The problem is discussed in App. II of this report.

A P P E N D I X I

CONTACT POTENTIAL VARIATION IN A LOG-LINEAR D. C. AMPLIFIER

OFFICIAL USE ONLY

PROJECT REPORT #6

ASMC-I

XAP-95185-BX

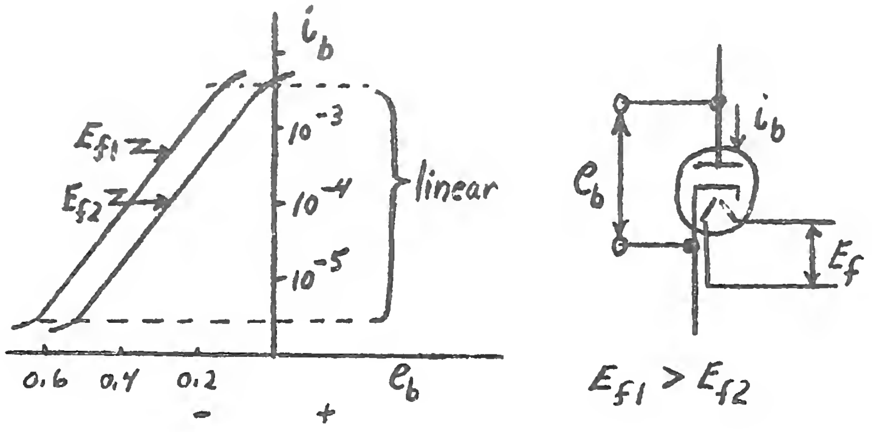
CONTACT POTENTIAL VARIATION
IN A LOG -- LINEAR D.C. AMPLIFIER -- AN INVESTIGATION

Date: 2-15-52

Lt. Comdr. Clarke

CONTACT POTENTIAL VARIATION
II. A LOG -- LINEAR D. C. AMPLIFIER -- AN INVESTIGATION

It is well known that if a diode is operated within certain ranges of filament current, the plate to cathode potential will be (over a given range) a logarithmic function of the plate current.



The slope of the log -- linear plot, within a relatively wide range of E_f, is not a function of E_f but the magnitude of e_b for a given i_b varies widely with E_f for a typical diode. Thus if one is interested in rate (de_b/di_b), relatively slow changes of E_f are insignificant in effects on output. However, if the interest is in current level (absolute values), then the significance of E_f is great.

The problem investigated was that of nullifying the effect of varying E_f on current level measurement in the range 10^{-10} through 10^{-4} ampere.

A review was made of the fundamental theory[1] behind this behavior of the diode. Basically the variation is due to change in electron affinity of the cathode with changes in cathode temperature. If an electron requires an energy

$$1/2\ mv_{min}^2 = W \qquad 1$$

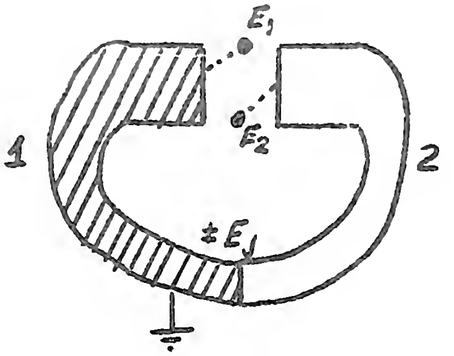
to escape from the surface of the cathode and we express this as

$$\Phi e = W \qquad 2$$

where e is the electronic change and Φ the potential corresponding to the kinetic energy in 1, then we call Φ the electron affinity and W the work function. Now the work function has been shown to be

$W = W_o + 3/2\ KT$ where
K = Boltzmann's constant
T = Temperature degrees Kelvin
W_o = Work function value at absolute zero

The relation of the work function to contact potential for two metals close together but not metallically connected is shown.



If metal 1 is grounded and an electron just escapes the surface its kinetic energy is converted to a potential energy $\Phi_1 e = W_1$ and has a potential with respect to ground of

$E_1 = -\Phi_1 = -W_1/e$. Similarly for E_2 except for the very small junction potential E_j.

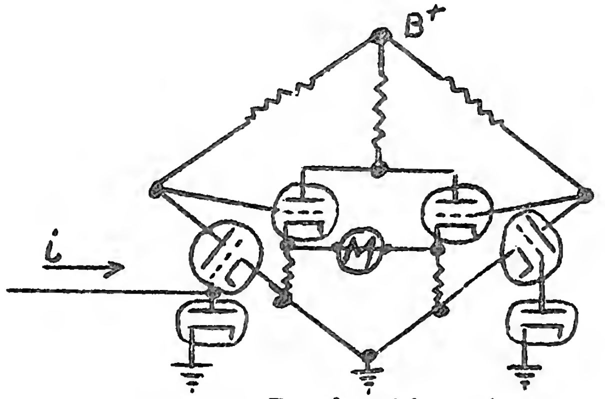
$$E_2 = -\Phi_2 \pm E_j = -W_2/e \pm E_j$$

Now the difference of potential (contact potential) is

$$E_2 - E_1 = \Phi_1 - \Phi_2 \pm E_j.$$

E_j is negligible compared to $(E_2 - E_1)$ in common diodes. Assuming the temperature of metal 1 held constant, it is seen that the measured contact potential is an inverse function of the temperature of metal 2 if the metal 1 has a greater electron affinity when both are at the same temperature (the case in common diodes).

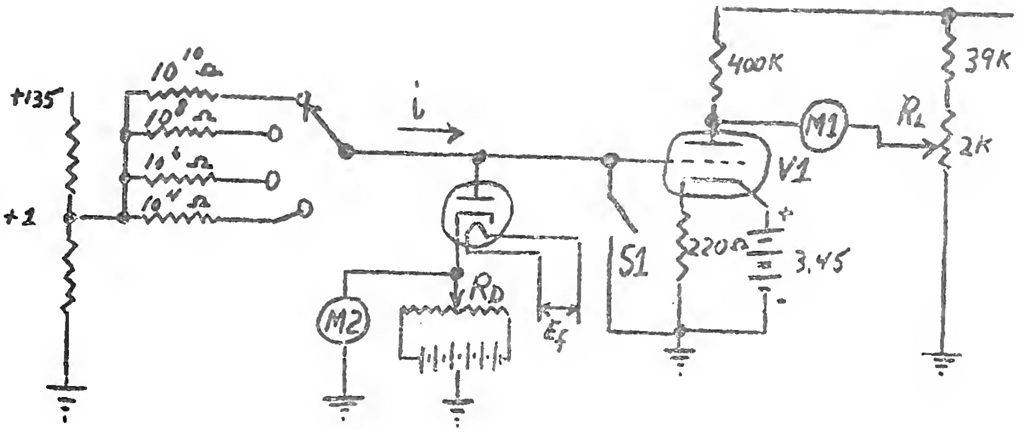
Inspection of the basic circuit used in this investigation[2] and consideration of the remark that instability arises from diode drift gave rise to the assumption that it was the purpose of this investigation to verify. The assumption was that if, in consideration of the basic theory above, there could be found a common cathode duo-diode then the differential output potential would not be a function of contact potential variation.



Basic Circuit

[illegible] and measurement of some stock diodes [illegible] check of tube handbooks revealed that the common cathode [illegible] was not available (6AN6 not in stock). The decision was made to use the diode sections of some duo-diode triodes (diode connected). The selected types were 6BF6, 6AT6, and 6J6.

The measurement circuit used is shown.



The measurement procedure was

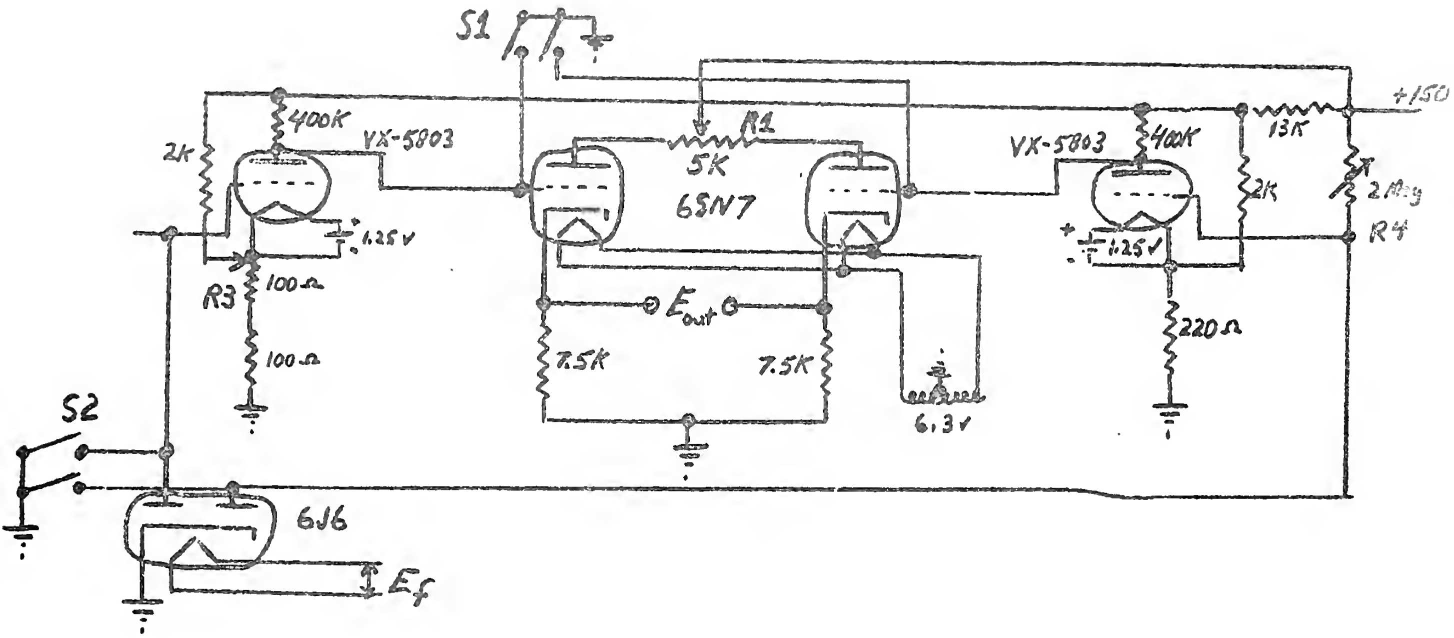
1. Closed S1 and varied R_L until M1 read zero. This established a balance for the grid of V1 grounded.

2. Opened S1 and varied R_D until M1 again read zero (for each current selection). This reestablished the grid of V1 at ground.

3. Read the voltage M2. Since $e_b + M_2 = 0$, then the current through the diode is that established by the selector switch and the input voltage and $M2 = -e_b$.

These measurements were repeated for various values of heater voltage. (Note the assumption that the grid current of the electrometer tube, V1, is negligible. This current is less than 10^{-14} ampere for tube operating conditions as set up). The results are plotted in Figures 1, 2, 6 and 7. Inspection of the plots of Figures 1 and 2 indicated the 6BF6 and 6AT6 are none too linear over the range considered. The plot of figure 3 showed the slope varies considerably for different current ranges for the 6BF6 whereas the plot of Figure 4 showed the slope to be very nearly the same at different current ranges in the vicinity (±0.5 volt) of $E_f = 4$ for the 6AT6. Inspection of Figure 6 showed the very good linearity of the 6J6 while Figure 5 showed the slope constancy desired at varying values of i_b in the region of $E_f = 5.5$. It was decided to use the 6J6 for the remainder of the investigation.

The remainder of the investigation was conducted using the following circuitry.



The procedure for measuring E_{out} for the various input currents at the selected values of E_f was:

1. With S_1 closed R_1 was varied to give $E_{out} = 0$.

2. With S_1 open and S_2 closed R_3 was varied to give $E_{out} = 0$.

3. With S_2 open R_4 was varied to give $E_{out} = 0$ for $i = 10^{-4}$ amps.

4. The various other currents were then fed to the input and E_{out} recorded.

The above measurements were repeated for various values of E_f. Also the measurements were made with the circuit zeroed for different values of E_f. The results are plotted in Fig.'s. 8, 9 and 10. Fig. 8 was made from data recorded with R_4 out of the circuit in order to observe circuit stability without a fixed reference current through the reference diode. These curves indicate that the reference diode serves as a fair potential "standard" without a given current through it especially in the range 10^{-4} through 10^{-3} ampere input for E_f between 5.5 and 6.0. The curves for zero-set at E_f = 5.5 (Fig. 9) were chosen as best since their slope is least in the region of E_f between 5.0 and 6.0 volts. The failure of the output slope at 10^{-10} amperes to be the same as that at 10^{-3} through 10^{-4} amperes was to be expected from observation of the wide divergence of diode characteristics as shown on Fig. 5 at this current value. (It is emphasized here that the investigation pursued was designed not to check differential

output against the filament voltage variation of the common cathode duo-diode in comparison to two matched diodes. The purpose was to measure per se the common cathode duo-diode differential output through various heater voltage ranges and current ranges to verify the assumption that contact potential effects would cancel. Note that contact potential variations in the matched diodes need not be in the same amount or the same direction.)

The errors are tabulated for E_f = (5.5 ±0.5) volt

I

i_b ↓ \ Zero Set E_f →	5.5	5.0	4.5	$\Delta e_b/\Delta E_f$
10^{-3}	0.00	-0.02	-0.02	
10^{-6}	0.03	-0.04	-0.04	
10^{-4}	0.00	-0.05	-0.07	

II

i_b ↓ \ Zero Set E_f →	5.5	5.0	4.5	Error Volts out/fil. Volt/decade
10^{-3} to 10^{-6}	-0.015	0.02	0.02	
10^{-6} to 10^{-4}	+0.015	+0.005	+0.015	

Since the voltage gain of the circuit was 1.06, the error is equivalent millivolts input per filament volt change per decade current was

III

i_b ↓ \ Zero Set E_f →	5.5	5.0	4.5	Error MV in/fil.V/decade
10^{-3} to 10^{-6}	-14.1	+18.8	+18.8	
10^{-6} to 10^{-4}	+14.1	+14.7	+14.1	

As a comparison to the possible error when using matched diodes in separate envelopes, the median possibility of error wherein one diode contact potential remains constant while the other varies (assuming the characteristics of Figure 5) was deduced.

IV

i_b ↓ \ Zero Set E_f →	5.5	$\Delta e_b/\Delta E_f$
10^{-8}	-0.31	
10^{-6}	-0.27	
10^{-4}	-0.26	

V

i_b range	error in volts out/fil. volt/decade
10^{-3} to 10^{-6}	0.145
10^{-6} to 10^{-4}	0.133

VI

i_b range	error in eq. millivolts in/fil. volt/decade
10^{-3} to 10^{-6}	137.0
10^{-6} to 10^{-4}	125.0

Comparison of table VI and column one of Table III showed that the possibility of error in the matched diode differential output is greater than that of the common cathode duo-diode by a factor of 125/14.1 = 8.85 for the current range of 10^{-6} to 10^{-4} ampere and 137/14.1 = 9.71 for the current range of 10^{-8} to 10^{-6} ampere in the case of this particular set of characteristics.

The conclusion derived from this investigation was that a common cathode duo-diode is indicated as a good possibility for stabilizing differential inputs of the log-linear type against contact potentail variations. The experimental results (in all cases investigated) point up the fact that such contact potential changes as did occur were in the same direction and simultaneous, thus bearing out the original assumption. It is to be noted that, although the results obtained in this investigation were far from hoped-for perfection in stabilization, near perfect stabilization by use of a diode designed to satisfy the circuit requirements offers great possibilities.

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1. Theory of Thermionic Vacuum Tubes, E. L. Chaffee, McGraw-Hill Book Co., Inc. 1933 p. 62.

2. Westinghouse Logarithmic Amplifier, R. L. Ramp, WAPD, Symposium on Reactor Instrumentation, WAPD 21 June 5-6, 1950, Part II.

MODEL DATE 24 Ja. 1952

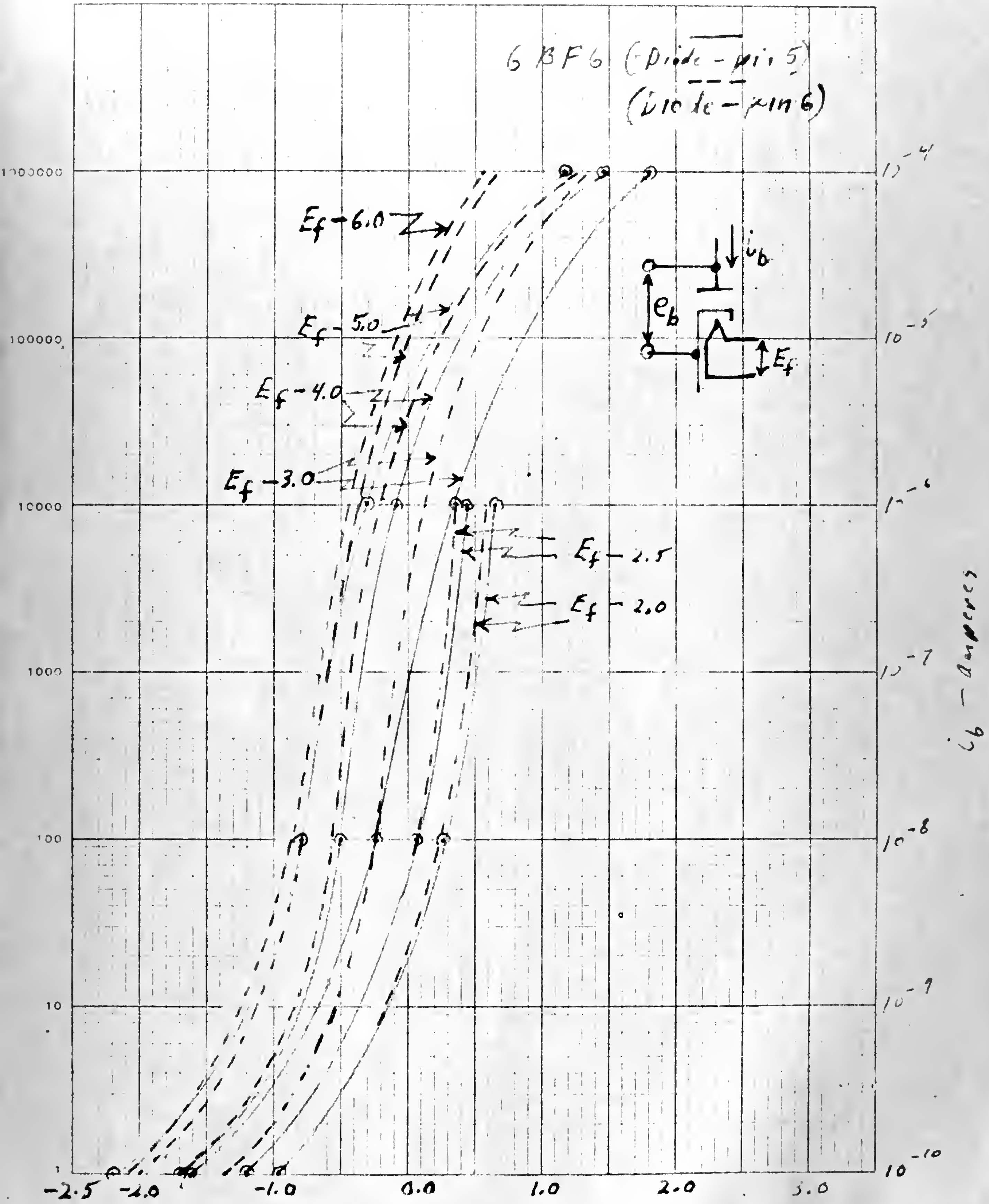


Fig. 2

MODEL DATE 31 Jan 1952

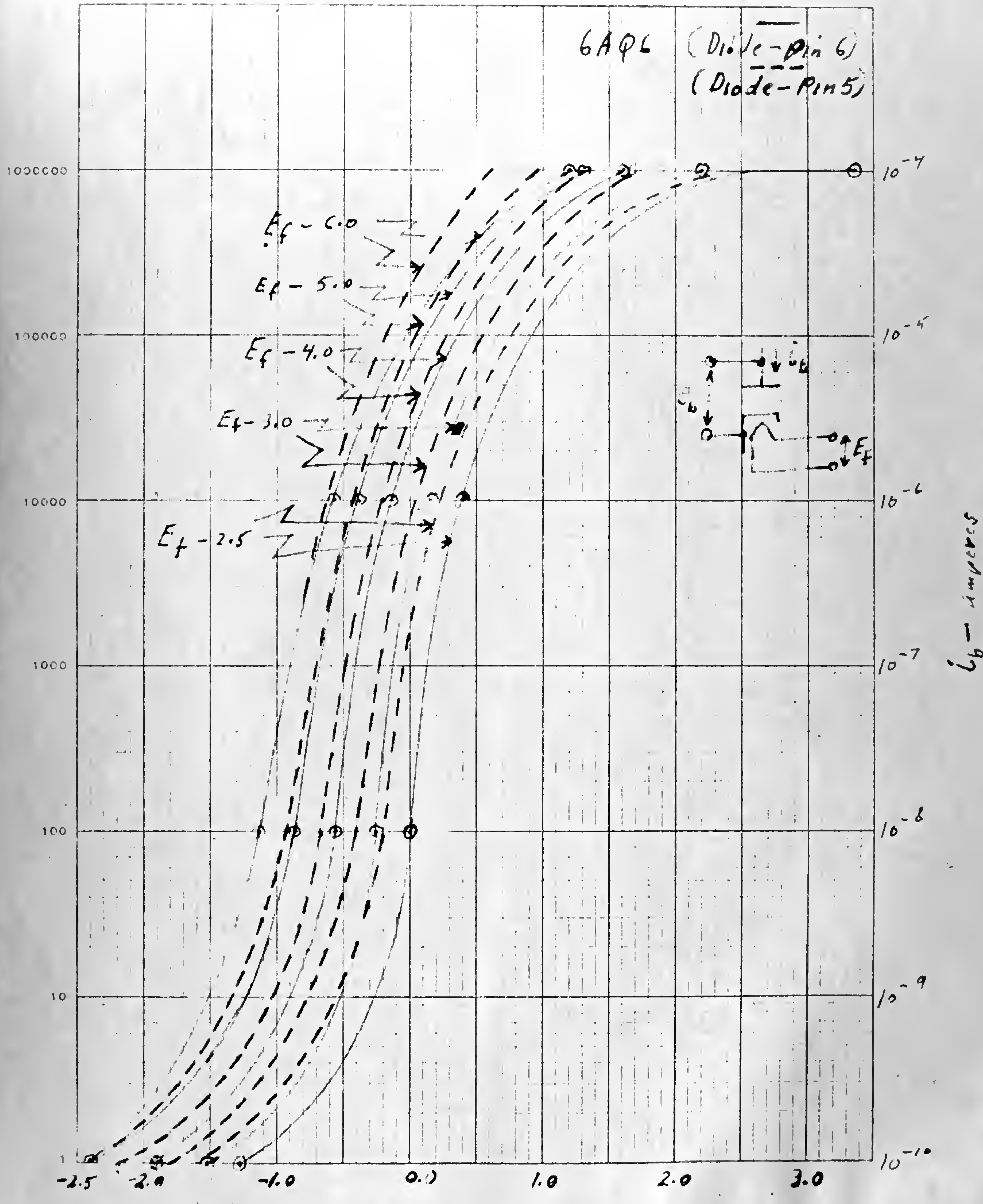


Fig.2 e_b - volts

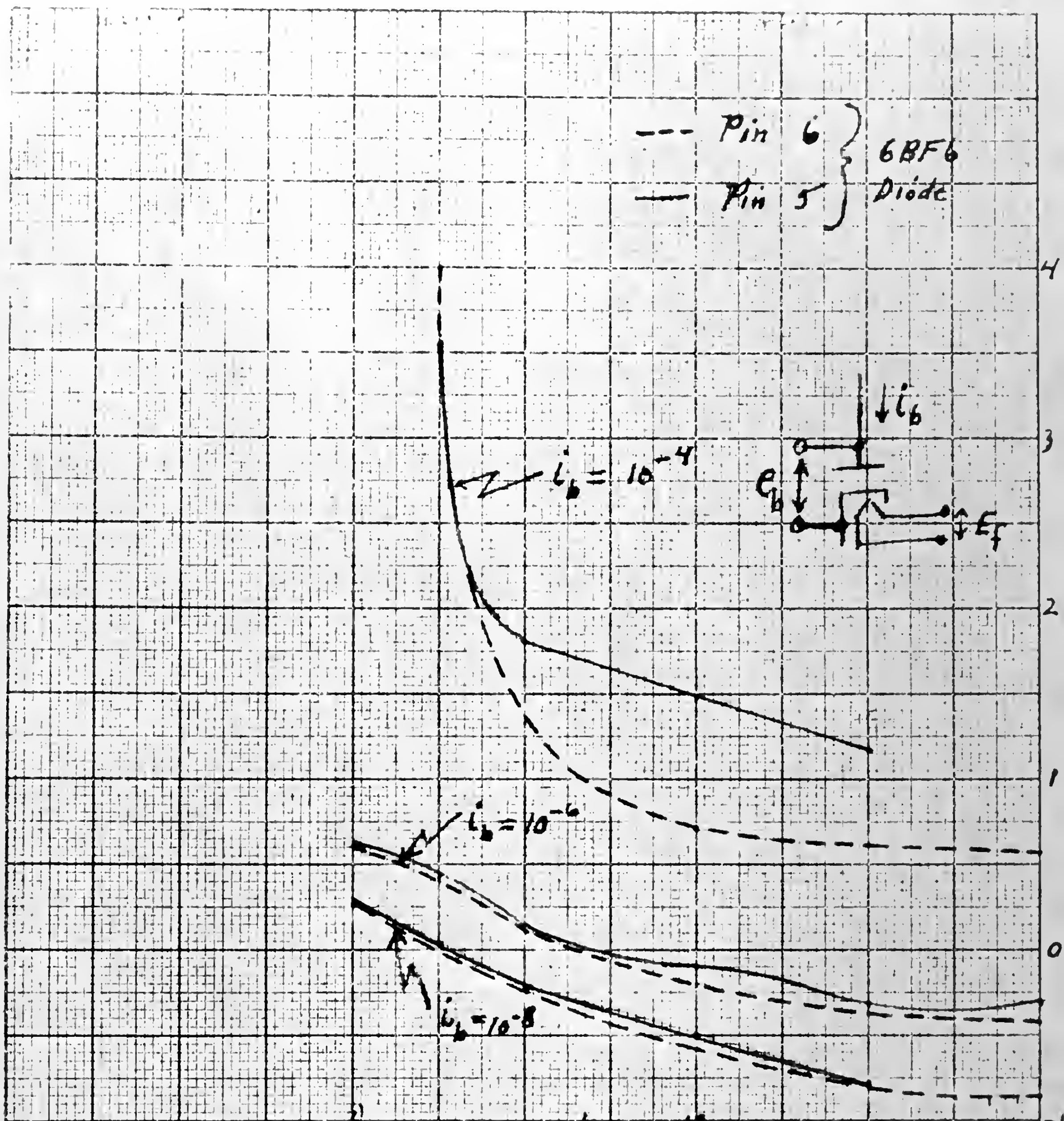

--- Pin 6
— Pin 5
6BF6 Diode
i_b = 10^-4
i_b = 10^-6
i_b = 10^-8
i_b = 10^-10
i_b
e_b
E_f
e_b vs E_f (i_b = constant)
0
1
2
3
4
5
6
-1
E_f - Volts

Fig.3

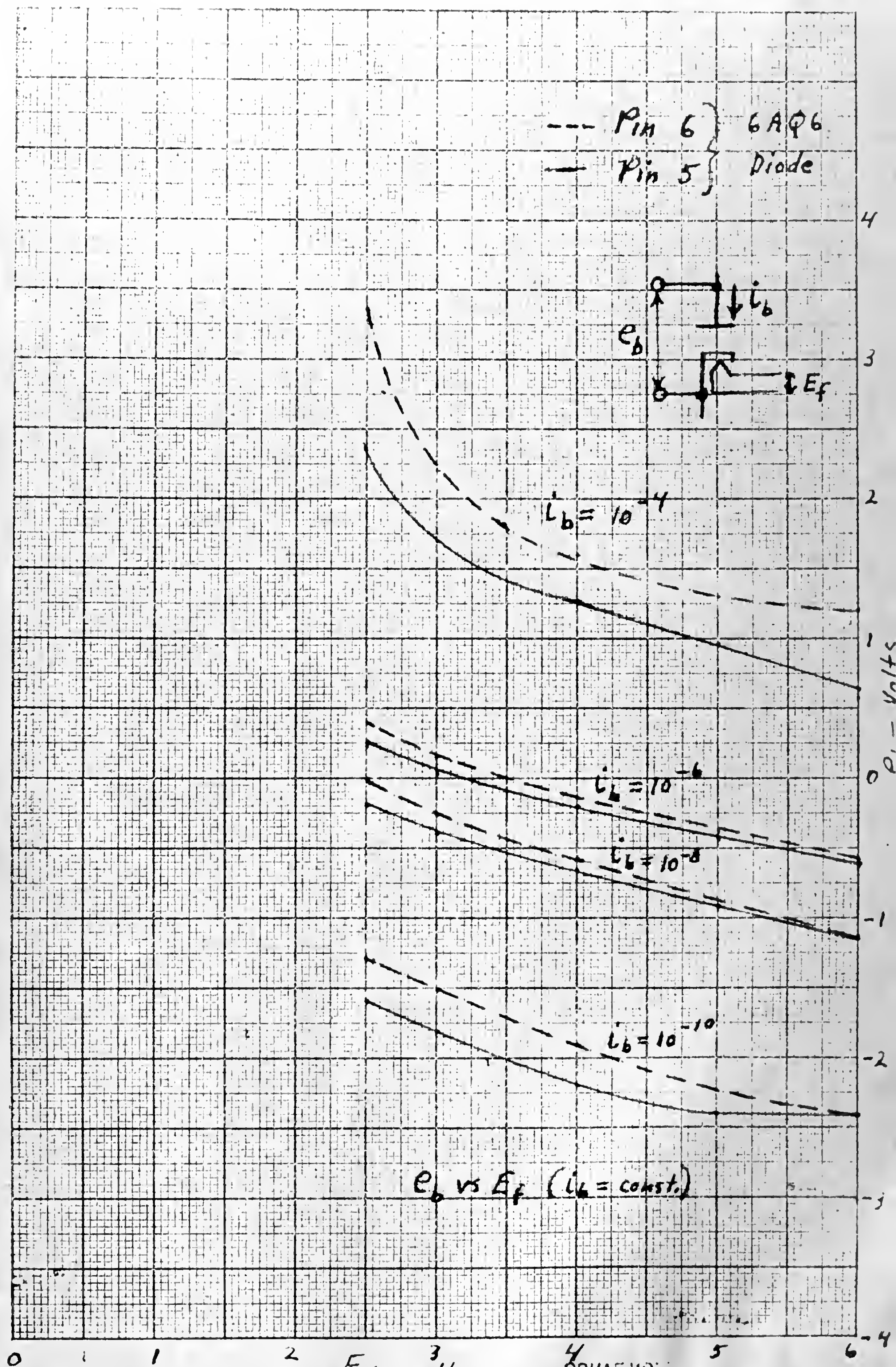
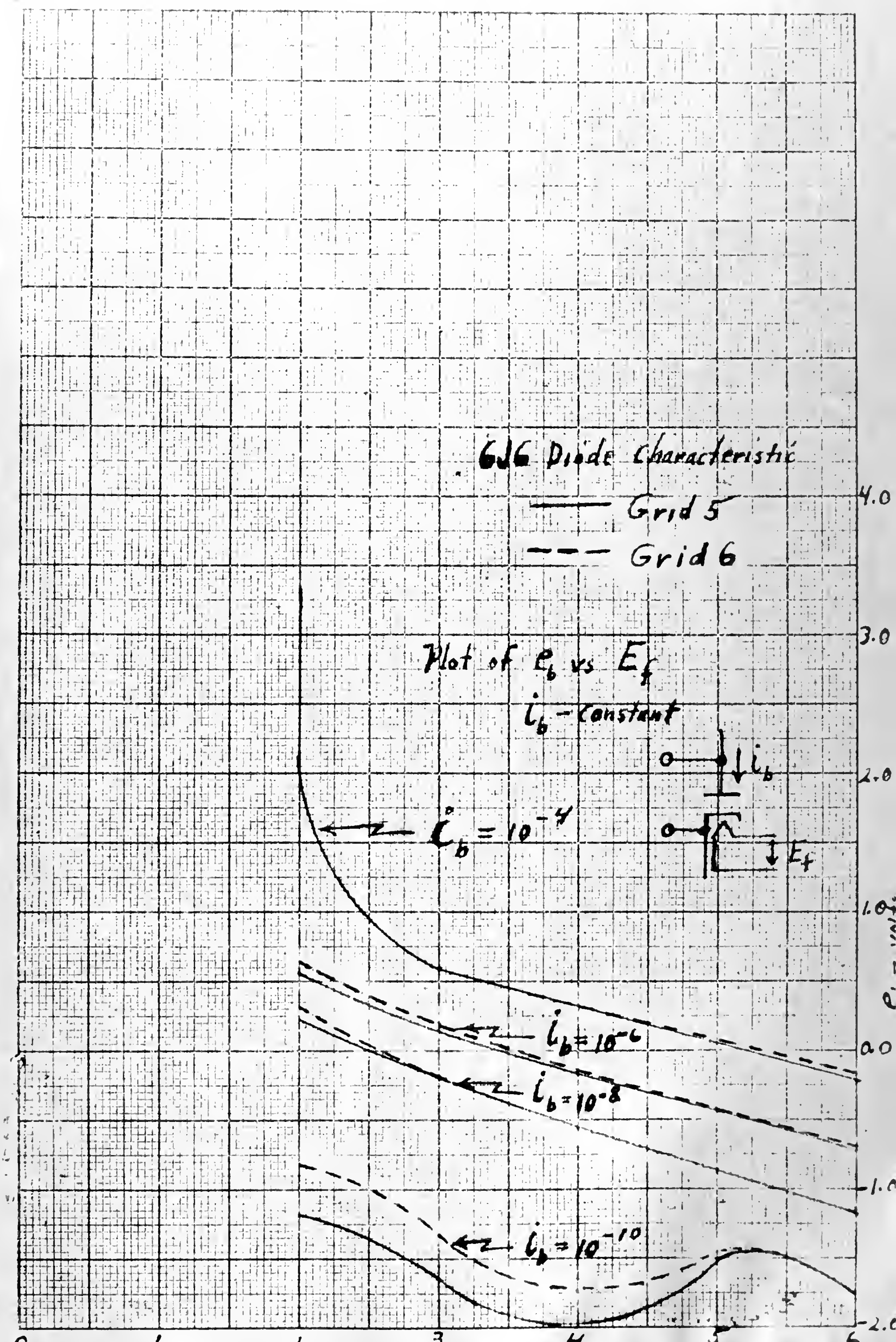

--- Pin 6
— Pin 5
6AQ6 Diode
i_b
e_b
E_f
$i_b = 10^{-4}$
$i_b = 10^{-6}$
$i_b = 10^{-8}$
$i_b = 10^{-10}$
e_b vs E_f (i_b = const.)
e_b - volts
4
3
2
1
0
-1
-2
-3
-4
0
1
2
3
4
5
6
E_f - volts

Fig. 4



6J6 Diode Characteristic
Grid 5
Grid 6
Plot of e_b vs E_f
i_b - Constant
i_b
E_f
$i_b = 10^{-4}$
$i_b = 10^{-6}$
$i_b = 10^{-8}$
$i_b = 10^{-10}$
4.0
3.0
2.0
1.0
0.0
-1.0
-2.0
0
1
2
3
4
5
6
E_f - Volts

Fig 5

MODEL DATE 4 Feb 1952

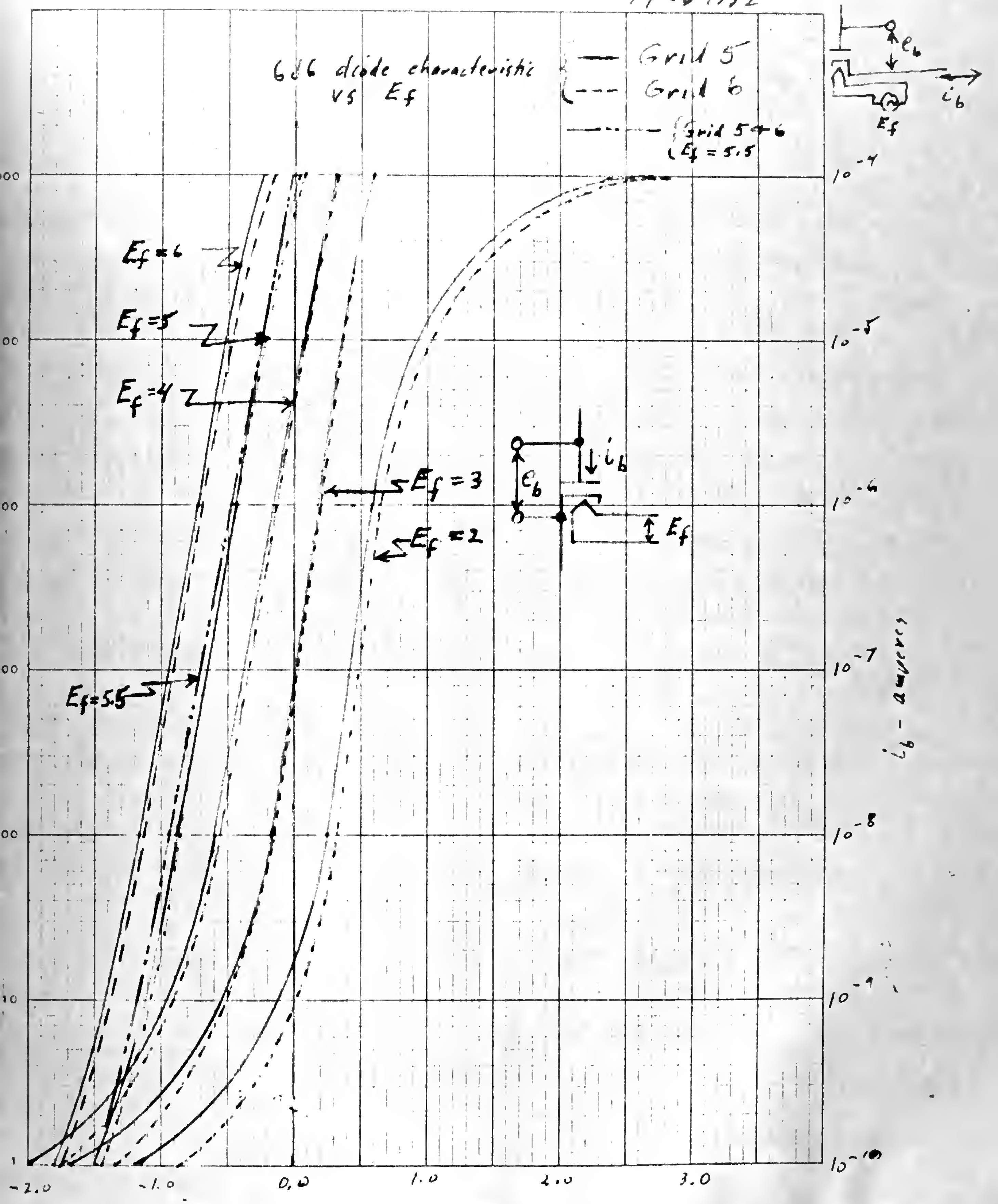


Fig. 6 e_b - volts

MODEL DATE

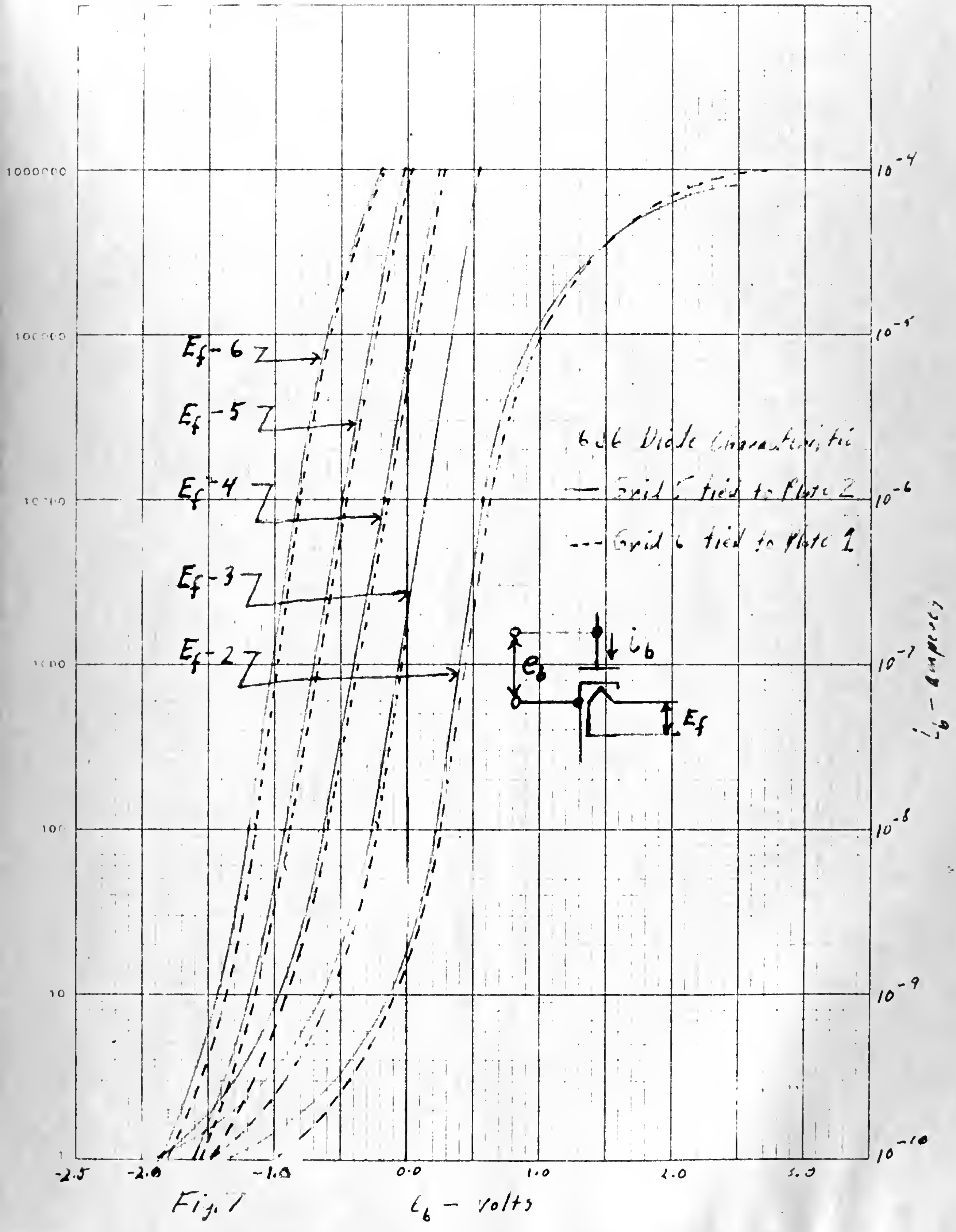


Fig. 7

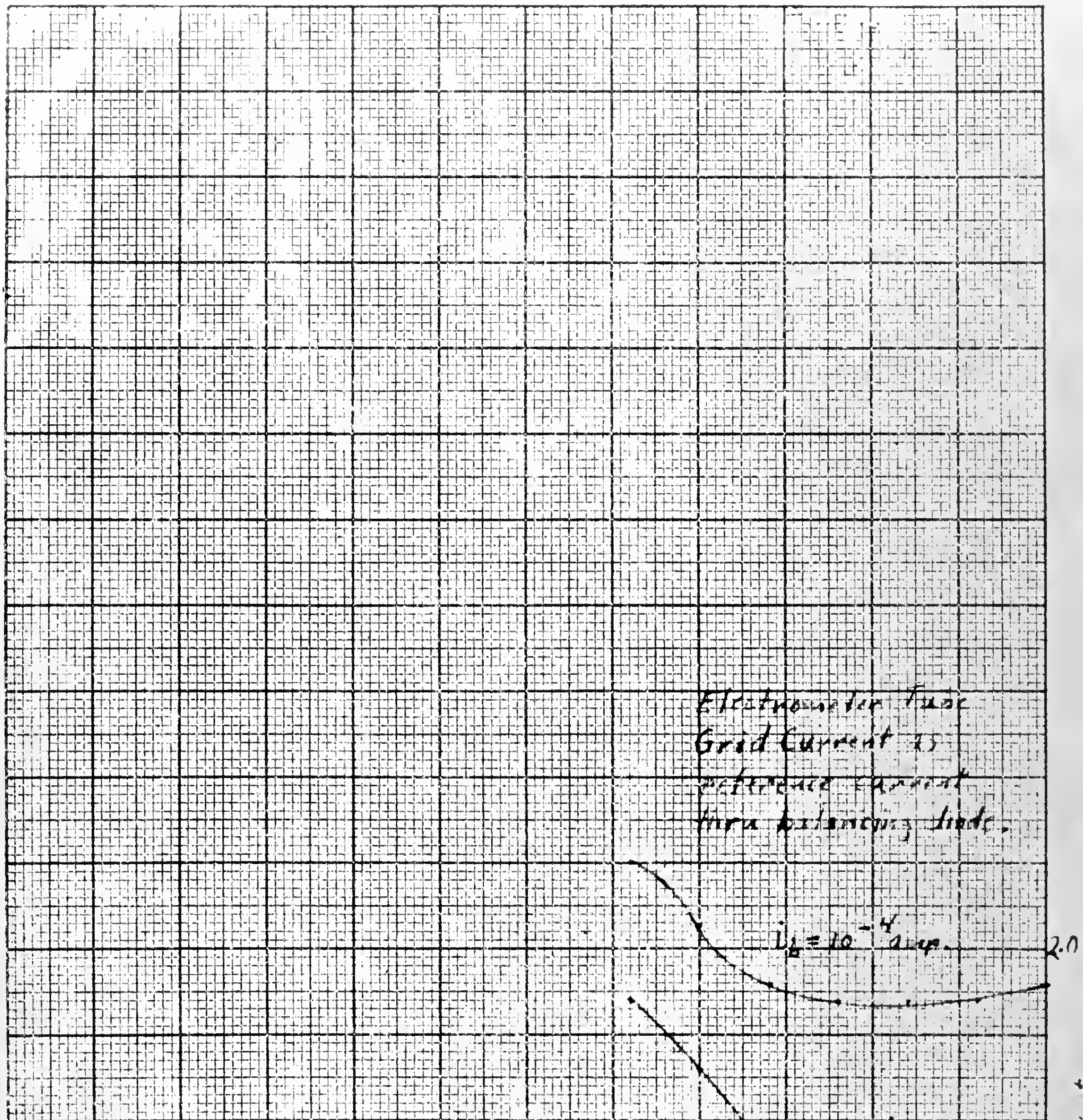

Electrometer Tube
Grid Current vs
reference current
thru balancing diode.
$i_b = 10^{-4}$ amp.
$i_b = 10^{-6}$ amp.
$i_b = 10^{-8}$ amp.
$i_b = 10^{-10}$ amp.
2.0
1.0
0.0
4.0
5.0
6.0
E_f - Volts

Fig. 8

WESTINGHOUSE ELECTRIC CORPORATION

CURVE NO.

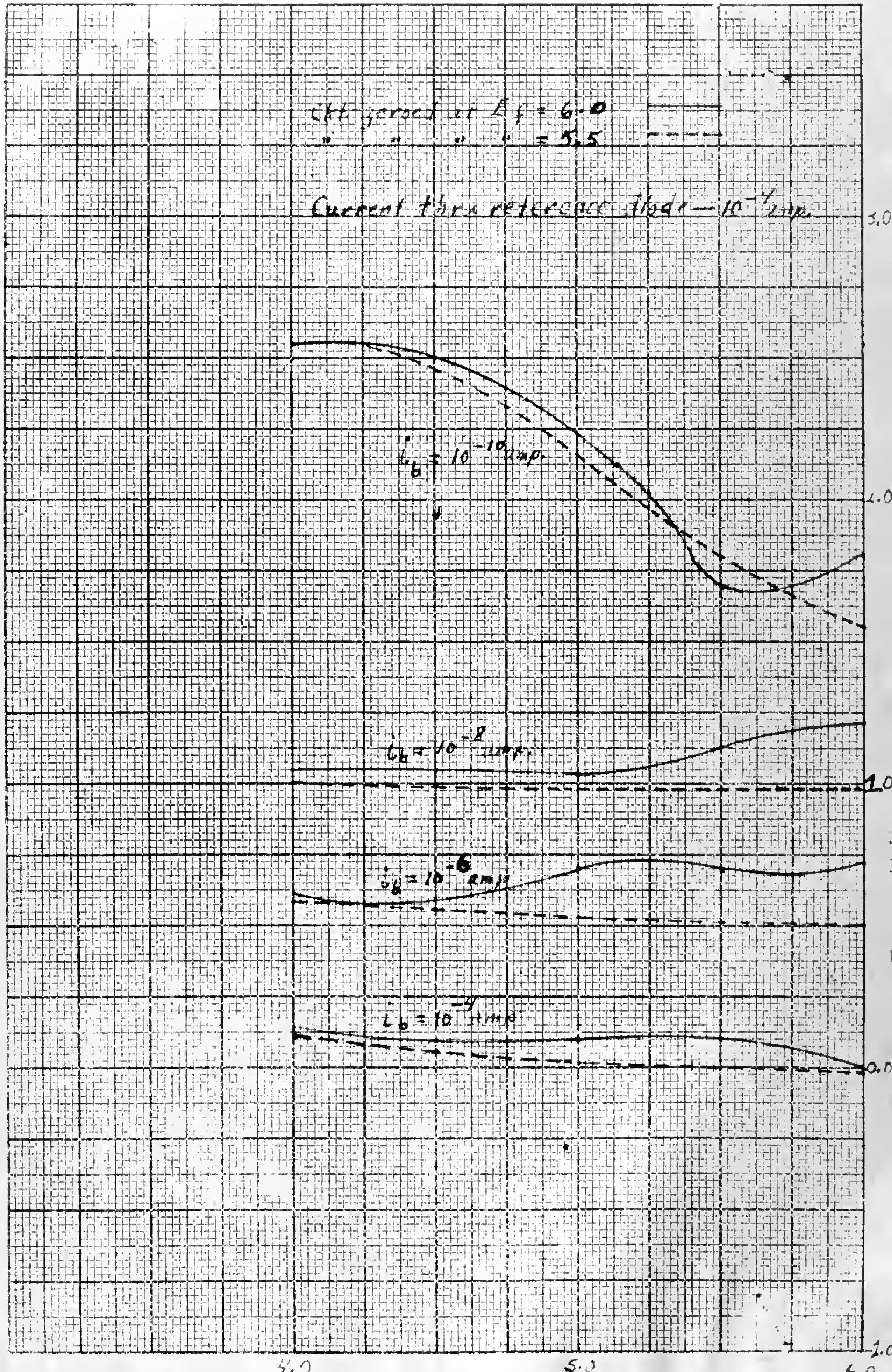


Fig. 1

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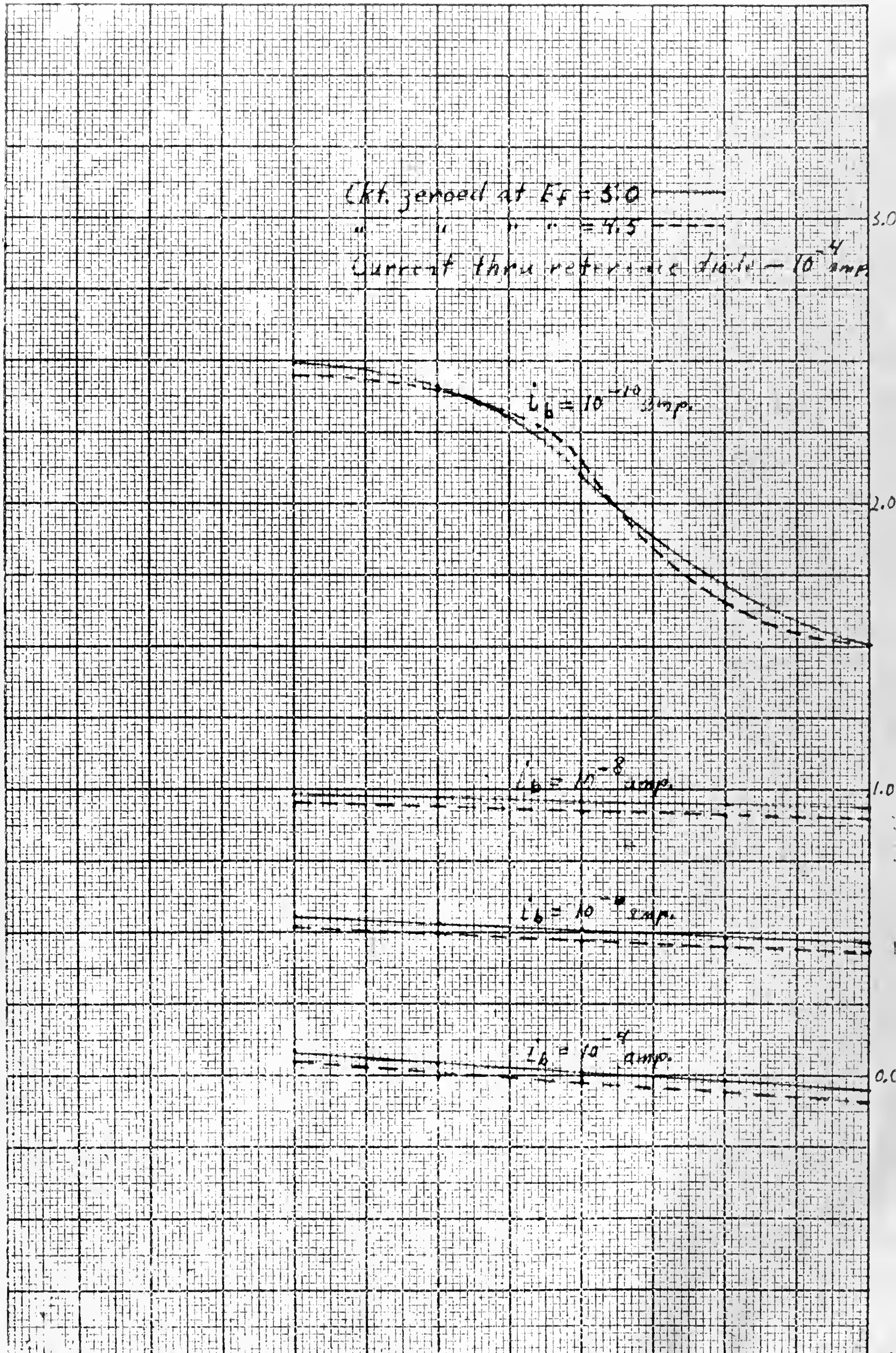


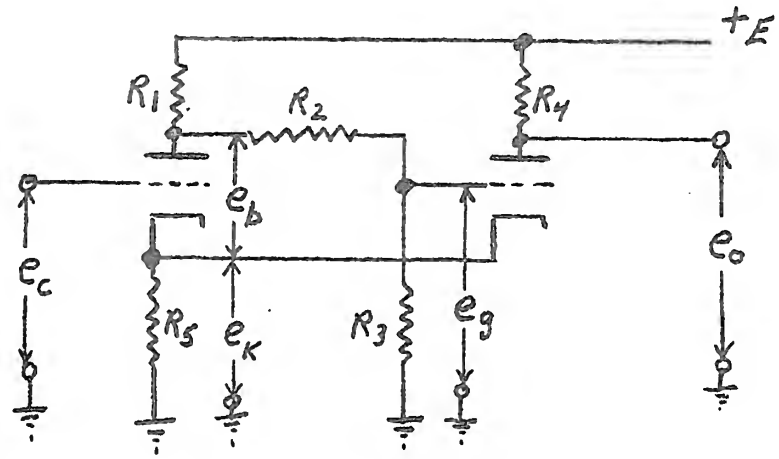
Fig. 10 E_f – Volts

A P P E N D I X II

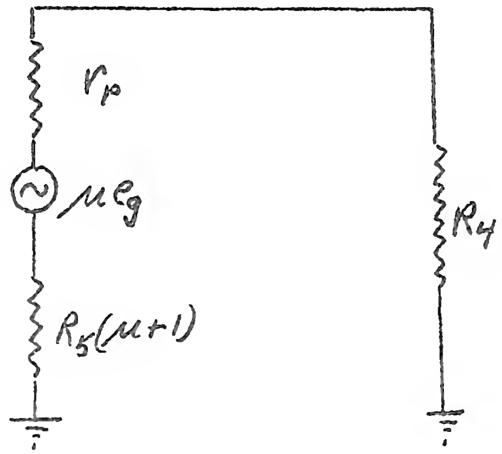
THE SCHMIDT TRIGGER CIRCUIT

THE SCHMIDT TRIGGER CIRCUIT

The occasion for investigation of the behaviour of this circuit was brought out in the narrative. The basic Schmidt trigger, a positive feedback amplifier, having two stable states is shown:



In the absence of input bias V_2 is "On". We have the elementary equivalent circuit.



Where $e_g = \frac{R_3}{R_1 + R_2 + R_3} \times E$

The interest here was to derive mathematically an expression giving the relationship between the various parameters and the hysteresis. In the above circuit with V_2 conducting, if e_c is raised to a certain level, V_1 begins to conduct. This conduction is amplified by V_2 and fed back positively to cause a rapid switching. If then e_g is reduced below this "trip" level a given amount V_2 conducts again and the circuit is "reset" to its previous state. This differential voltage, trip to reset, is termed hysteresis.

Notation:

e_{co} -- cut-off grid bias

E_{B1} -- Plate to cathode potential V_1

E_{B2} -- Plate to cathode potential V_2

e_K -- Cathode to ground potential

e_g -- Grid to ground potential of V_2

e_c -- Grid to ground potential of V_1

Subscript 1 -- V_2 Conducting

Subscript 2 -- V_1 Conducting

$$e_{K1} = \frac{\mu R_3 R_5 E}{(R_1+R_2+R_3)\left[r_p + R_5(\mu+1)+R_4\right]} \tag{1}$$

To trigger V_1 into conduction (Trip);

$$e_{c1} = i_{b1} R_K - e_{co} = e_{K1} - e_{co} \approx e_{K1} - \frac{E_{B1}}{\mu}$$

Substituting 1 for e_{K1}

$$e_{c1} = \frac{E}{\mu(R_1+R_2+R_3)}\left[\frac{\mu R_3 R_5(\mu+1)}{r_p+R_5(\mu+1)+R_4} - (R_2+R_3)\right] \tag{2}$$

To trigger V_2 into conduction (Reset)

$$e_g = e_{K2} - e_{co} = \frac{R_3}{R_2+R_3}(E - i_{b2}R_1) \approx e_{K2} - \frac{E-e_{K2}}{\mu} \tag{3}$$

$$e_{K2} = \frac{\frac{\mu R_3}{R_2+R_3}(E-i_{b2}R_1) + E}{\mu+1} \tag{4}$$

Now $i_{b2} = \dfrac{\mu\, e_{c2}}{r_p + R_5(\mu+1) + R_1}$

And $i_{b2}R_5 = e_{K2}$ or $i_{b2} = \dfrac{e_{K2}}{R_5} =$ (5)

$$= \frac{\mu\, e_{c2}}{r_p + R_5(\mu+1) + R_1} \tag{6}$$

$$\text{or } e_{c2} = \frac{e_{K2}\left[r_p + R_5(\mu+1) + R_1\right]}{\mu R_5} \tag{7}$$

Subst. 5 in 4

$$e_{K2} = \frac{E\left(\dfrac{\mu R_3}{R_2+R_3} + 1\right)}{(\mu+1) + \dfrac{\mu R_1 R_3}{R_5(R_2+R_3)}} \tag{8}$$

Subst. 8 in 7

$$e_{c2} = \frac{\dfrac{E\left(\dfrac{\mu R_3}{R_2+R_3} + 1\right)}{(\mu+1) + \dfrac{\mu R_1 R_3}{R_5(R_2+R_3)}}\left[r_p + R_5(\mu+1) + R_1\right]}{\mu R_5}$$

The hysteresis is

$h = e_{c1} - e_{c2}$

$$h = \left\{\frac{E}{\mu(R_1+R_2+R_3)}\left[\frac{\mu R_3 R_5(\mu+1)}{r_p + R_5(\mu+1) + R_1} - (R_2+R_3)\right]\right\}$$

$$-\frac{1}{\mu R_5}\left\{\frac{E\left(\dfrac{\mu R_3}{R_2+R_3} + 1\right)}{(\mu+1) + \dfrac{\mu R_1 R_3}{R_5(R_2+R_3)}}\left[r_p + R_5(\mu+1) + R_1\right]\right\} \tag{9}$$

Since R_2 and R_3 serve as a potential divider,

$$R_5 \ll R_2 + R_3 \gg R_1$$

$$R_3 \gg R_1 \ll R_2$$

$$R_3 \gg R_5 \ll R_2 \tag{10}$$

Using the approximations 10 in 9,

$$h = \frac{ER_3}{R_2+R_3} - \frac{E}{\mu} - \frac{E(R_2+R_3)\left[r_p+R_5(\mu+1)+R_1\right]}{\mu\left[(\mu+1)(R_2+R_3)\;R_5+\mu R_1R_3\right]}$$

$$- \frac{ER_3\left[r_p+R_5(\mu+1)\;+R_1\right]}{(\mu+1)(R_2+R_3)R_5+\mu R_1R_3} \qquad (11)$$

If $(\mu+1)(R_2+R_3)(R_5) \gg \mu R_1R_3$, which is true for commonly used circuits and a fair approximation for values wherein (12)

$$(\mu+1)(R_2+R_3)R_5 = 5\;(\mu R_1R_3)$$

Using 12 in 11,

$$h = \frac{E\;R_3}{R_2+R_3} - \frac{E}{\mu} - \frac{E}{\mu}\left\{\frac{r_p}{(\mu+1)R_5} + 1 + \frac{R_1}{(\mu+1)R_5}\right\}$$

$$- ER_3\left\{\frac{r_p}{(\mu+1)(R_2+R_3)R_5} + \frac{1}{R_2+R_3} + \frac{R_1}{(\mu+1)(R_2+R_3)R_5}\right\}$$

Since $\mu+1 \approx \mu$

$$h = -\frac{2E}{\mu} - \frac{E}{\mu^2R_5}(r_p+R_1) - \frac{E\;R_3}{\mu R_5(R_2+R_3)}(r_p+R_1) \qquad (13)$$

$$h = -\frac{E}{\mu}\left[2 + \frac{r_p+R_1)}{R_5}\left(\frac{1}{\mu}\quad\frac{R_3}{R_2+R_3}\right)\right]$$

From this we note that, considering R_5 the independent variable we have an equation in the form $y = A + B/X$, an hyperbola.

If R_1 be considered the independent variable we have $y = A + BX$, a straight line. By similar algebraic methods, the trip potential (e_{c1}) is

$$\frac{1}{R_5}\left(\frac{R_1}{2} - \frac{2}{5}\right)\frac{E}{2} \quad \text{if } R_5 \text{ is set so that } e_{K1} = \frac{E}{2}$$

With this background to aid, an experimental study was made of the circuit in production to arrive at design data for guidance in alteration to the optimum parameters to satisfy the requirements imposed for pulsed operation.

1. Small hysteresis.
2. Large output pulse amplitude.
3. Output pulse amplitude independent of input.

The circuit is as shown on figure 1. No attempt was made to analyze this circuit mathematically inasmuch as the convenient approximations for triode cut-off are not applicable. Further the ground point in this circuit is "floating" and the screen potential is a function of the bias voltage. The plotted data represent the work done and the effect on the circuit of the various parameters. It is to be noted that the variation of hysteresis with R_1 and R_5 (R_L and R_K respectively in the figure) is substantially of the form predicted for the triode circuitry.

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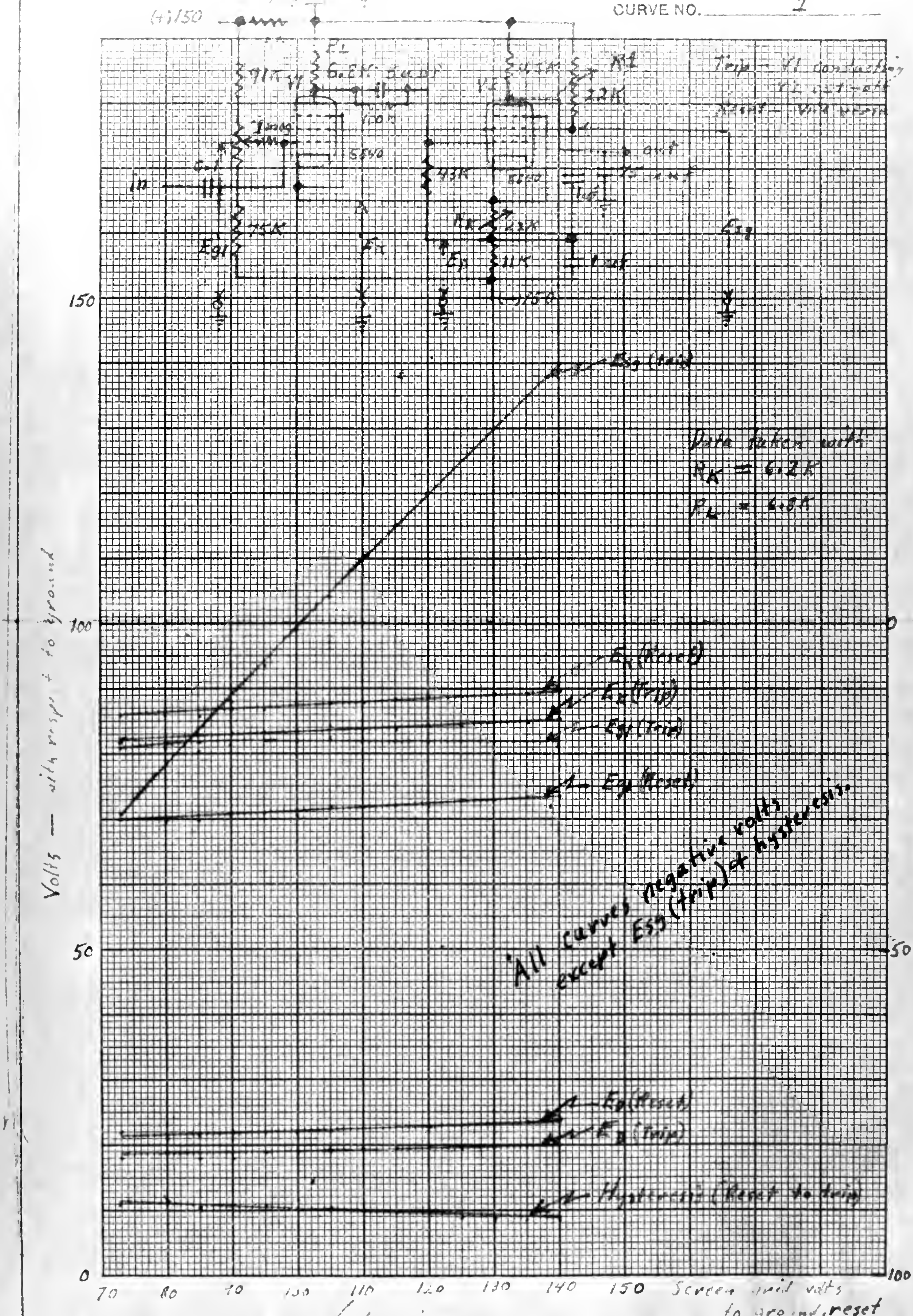

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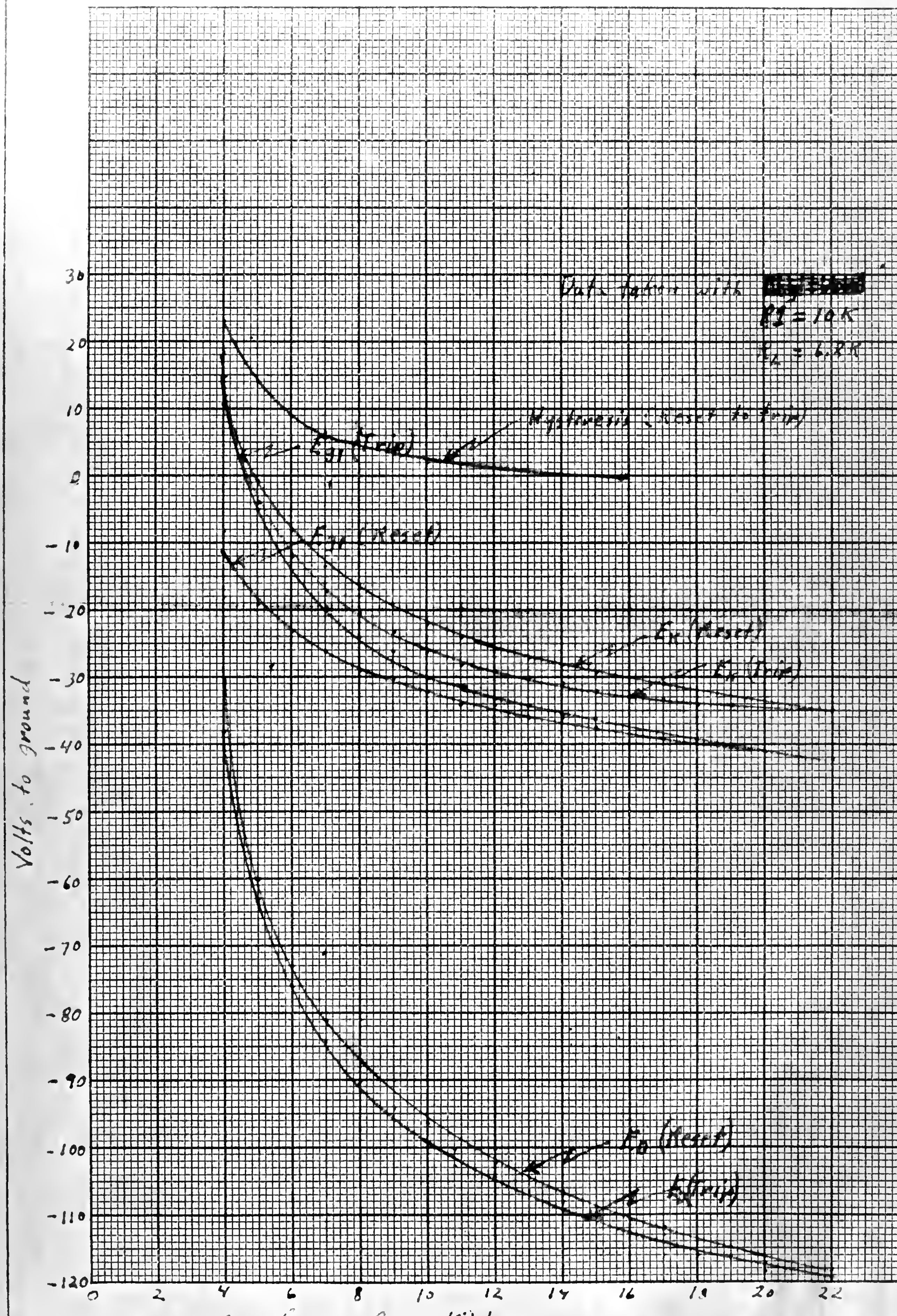

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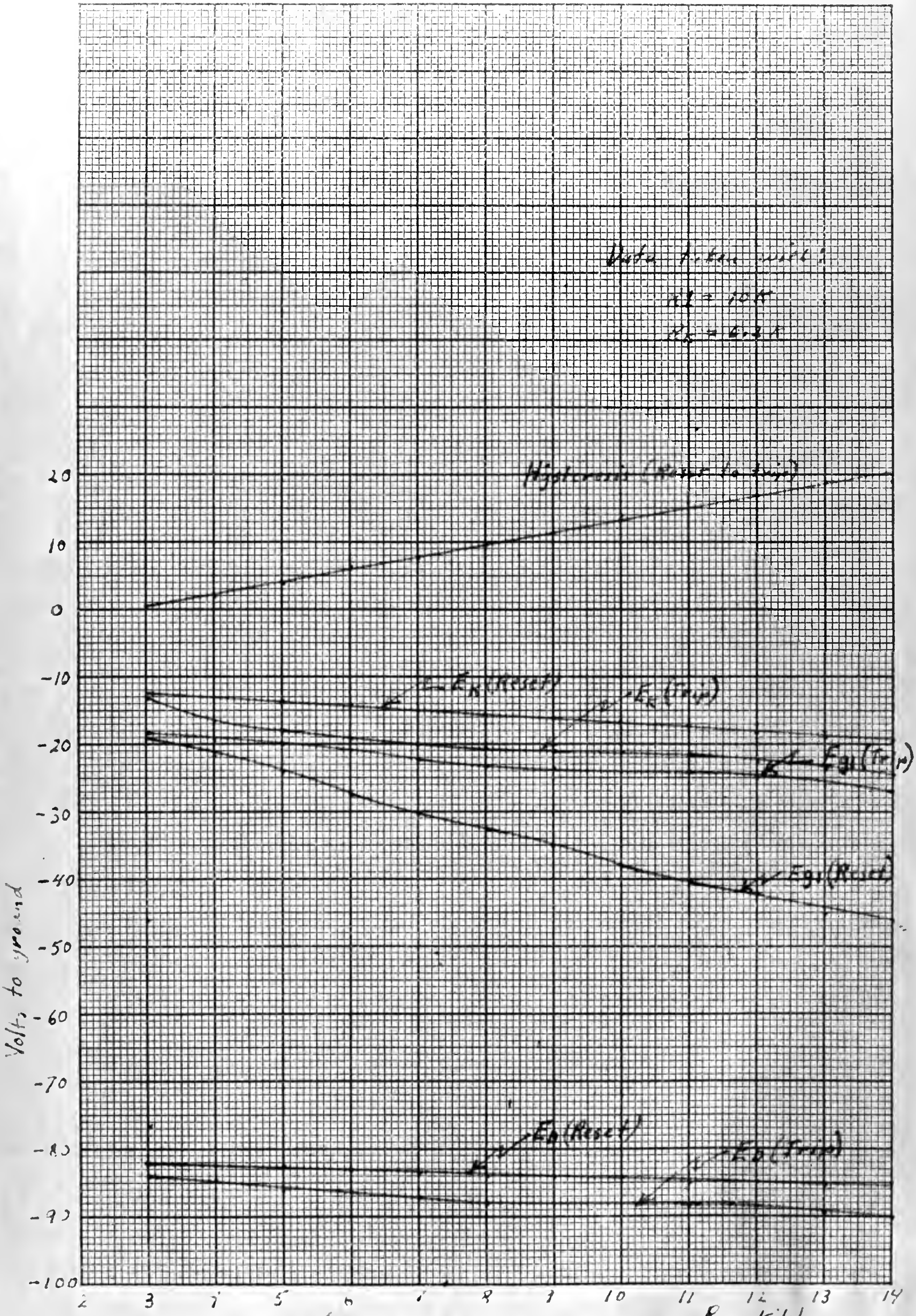

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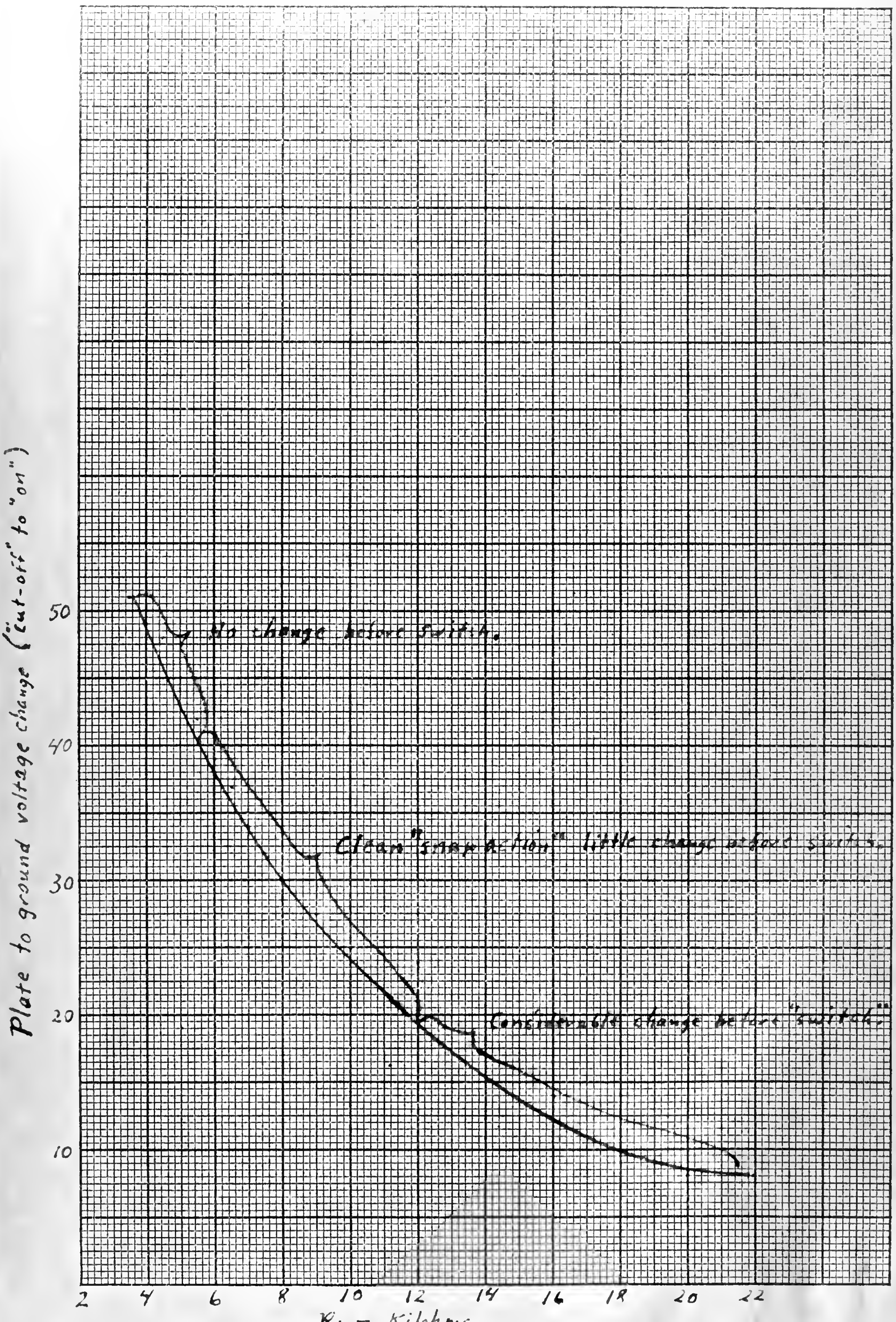
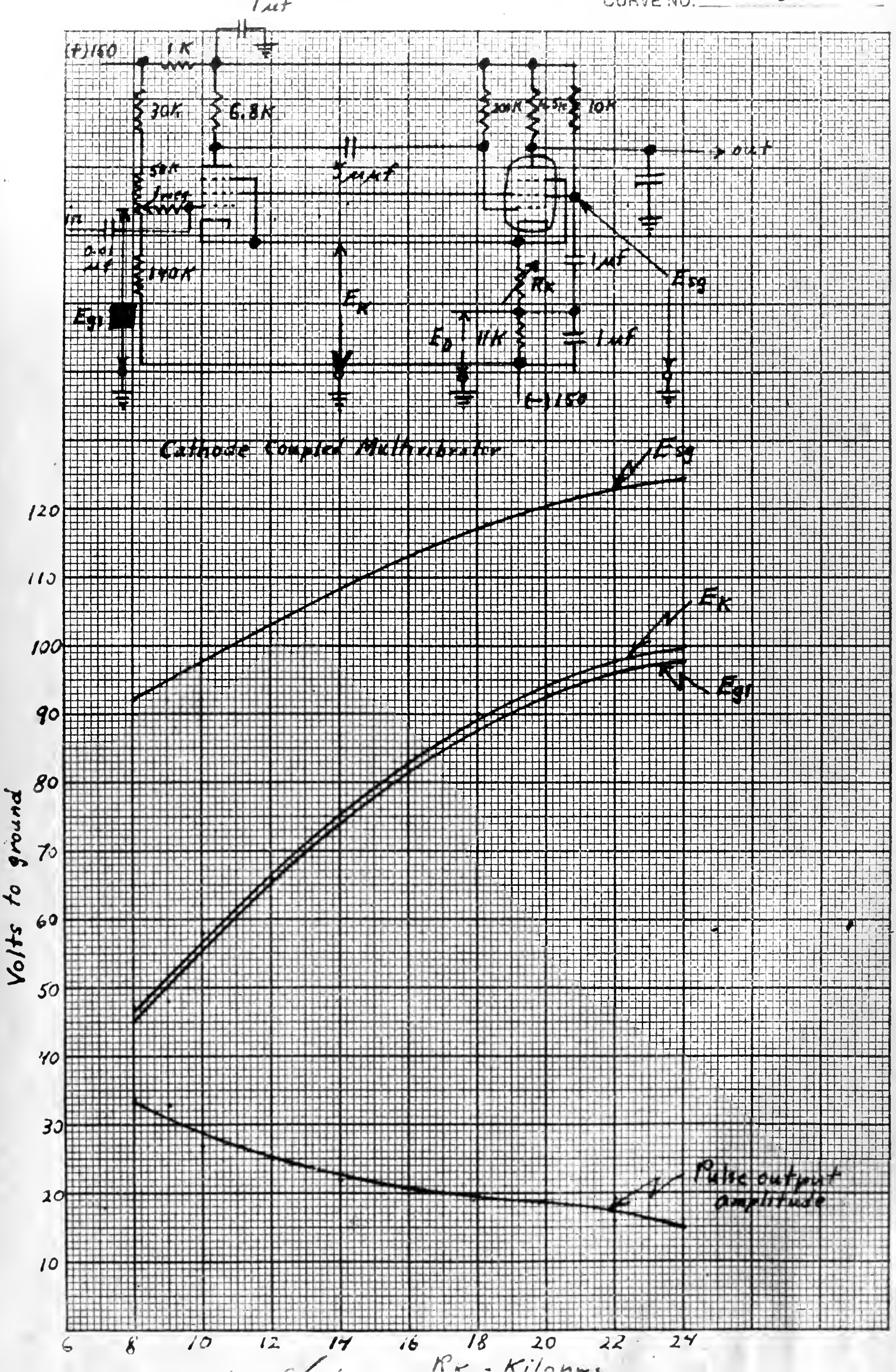


Plate to ground voltage change ("cut-off" to "on")
50
40
30
20
10
No change before switch.
Clean "snap action" little change before switch.
Considerable change before "switch."
2
4
6
8
10
12
14
16
18
20
22
R_K - Kilohms

WESTINGHOUSE ELECTRIC CORPORATION

CURVE NO. 5



A P P E N D I X III

THE DIRECT-COUPLED AMPLIFIER

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This digest of the literature relating to DC amplifiers, is made to permit selection and design of circuits with intelligent application of the criteria of gain, drift stability and band pass with due regard to circuit simplicity in accordance with present day knowledge.

A proper introduction to the DC amplifier problem is a discussion of the amplifier's basic drift limitations. Since, in such amplifiers, the output of one stage is directly coupled to the next, any change in DC level in the amplifier appears as an output signal. The possible changes causing this drift are:

1. Plate supply voltage change.
2. Component value change.
3. Cathode temperature variation.
   a. Contact potential, grid to cathode, change.
   b. Equivalent plate resistance change.
4. Tube aging with resultant change in R_p (emission change).

The drift voltages associated with plate supply changes are self-explanatory as are those due to component value changes. Those due to changes in emission and cathode temperature are more complex. Shown is a typical triode I_p - E_g curve for different heater voltages.

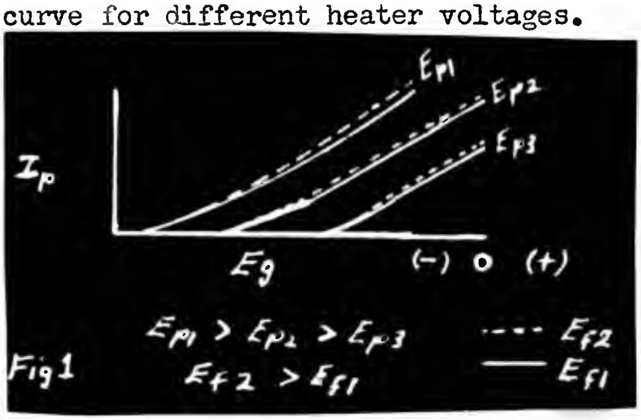

Fig. 1

If such a tube be operated near cut-off conditions it may be seen that there is a displacement of the curve corresponding to a grid voltage change and a change in slope corresponding to an effective change in r_p. But if the tube be operated in the region of linear characteristic, the curves are essentially parallel and the cnage with heater voltage corresponds to a grid-cathode potential change. Due to the direct coupling, plate-to-grid, the necessary potential level of cascaded stages is high with resultant high capacitance to ground and poor high frequency response in the basic DC amplifier. This necessary high potential with high shunt capacitance together with drift comprise the shortcomings of the basic amplifier. (Note: These drift considerations prevent use of a-c filament voltages in the basic amplifier.)

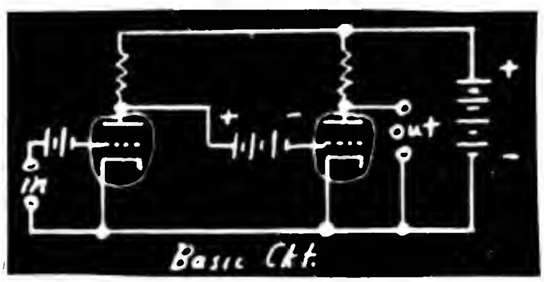

Fig. 2

The corrective measures in present day use to compensate for the above undesirable features of the DC amplifier are brought out in the remainder of this digest.

Direct-coupled amplifiers may be grouped to fall into the following classifications:

1. Simple cascaded.
2. Drift compensated -- simple.
3. Bridge balanced.
4. Cathode follower coupled.
5. Differential or current coupled.
6. Gas tube coupled.
7. Modulation systems (Chopper amplifiers)

Of the above classifications 7 is not discussed herein inasmuch as it may be more properly considered as an AC amplifier. The following approximation was used in the algebraic calculation of all circuit equations.

From $i_p = \left[\frac{e_p + \mu e_g}{r_p}\right]^n$ consider n = 1 and

$$r_p = \frac{e_p}{i_p}$$

we have $e_p = i_p r_p - \mu e_g$ $\qquad$ 1

In the linear region of the operating characteristic is where operation is assumed. It is valid in this region to assume the following equation when contact potential is treated (See fig. 1.).

$e_p = i_p r_p - \mu e_g + \mu V_f$ where V_f is a potential inserted 2 in series with the cathode.

Figures 3 through 8 cover in general the first two classification above. The mathematical expressions below the figures express the dependence of the output on the various parameters of the circuitry and supply voltages.

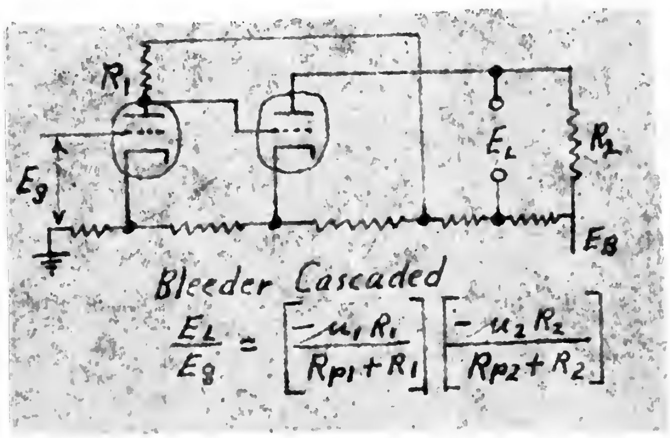

Fig. 3

Equation holds when the bleeder current is much greater than i_p. Not balanced for E_f , r_p or E_B . Cascade on common bleeder. Amplification varies with r_p.

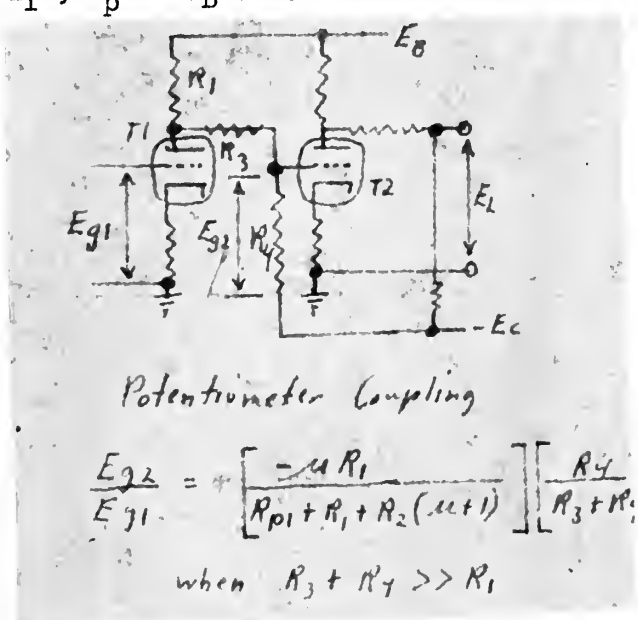

Fig. 4

High drift. Not balanced for E_f, r_p or E_B. Easily cascaded on common bleeder. Amplification varies with r_p.

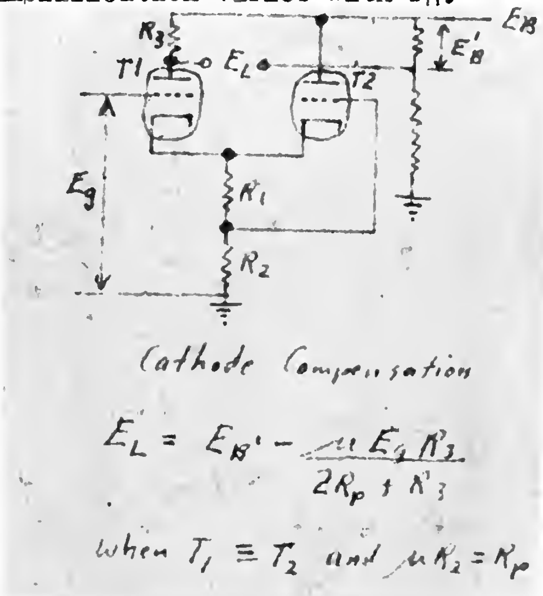

Fig. 5

Medium drift. Compensated for E_f but not for r_p or E_B. Can be cascaded on common bleeder. Amplification varies with r_p.

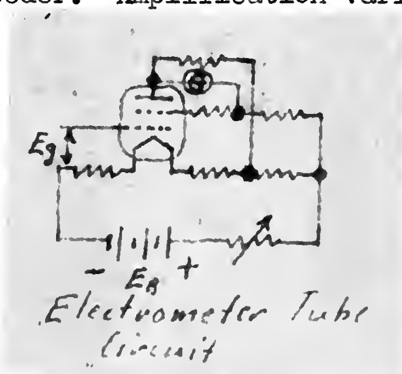

Fig. 6

Low drift. Accurate balance for E_B and r_p for short time use. Cannot be cascaded. Of interest for sensitive galvanometer deflection.

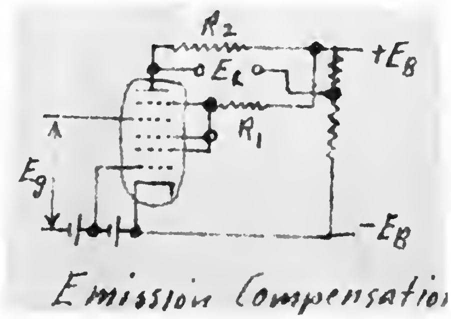

Fig. 7

Medium drift. R_1 adjusted so grids 1 and 4 have equal but opposite g_m. Compensated for E_f and E_{c1} but not for E_B and E_{c2}. Can be cascaded on common bleeder. Amplification varies with r_p.

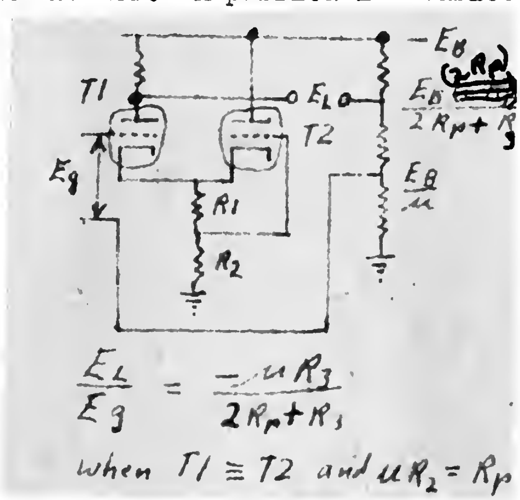

Fig. 8 Cathode and B supply compensation.

Low drift. Compensated for E_f and E_B but not for r_p. Can be cascaded on common bleeder. Amplification varies with r_p.

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Of the above circuits 3 and 4 are seldom used without compensation. They may be used as stabilized current cmplifiers by use of 100% feedback. Figure 5 is a typical compensated amplifier. It is stable with a stable supply voltage. The circuit of fig. 8 compensates for B supply variation by returning the input to a tap on the supply of value E_B/u. Figure 7 illustrates compensation for heater voltage and random emission effects. R_1 is adjusted to give grids 1 and 4 equal but opposite transconductance. Compensation includes all circuit elements common to both grids but not plate supply. Returning grid 4 to a tap on the bleeder across the plate supply corrects for changes in E_B similar to circuit fig. 8. Signals may be applied to either grid or as a differential amplifier to both.

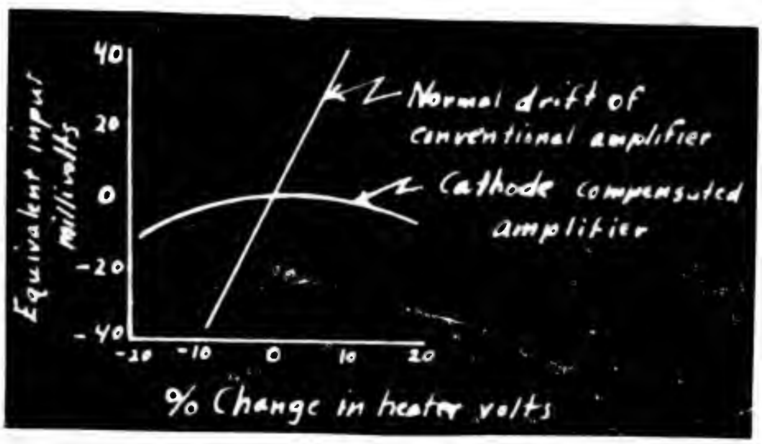


Fig. 9

Figure 9 is a plot of drift in a triode DC amplifier and in a cathode compensated amplifier. Circuit figure 6 can be compensated for short times by choice of circuit constants but difficulty in cascading limits its use to galvanometer circuits.

Figures 10 through 15 are of the bridge balanced type of DC amplifier.

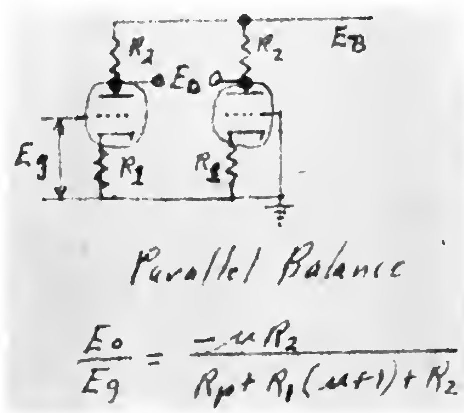


Fig. 10

Low drift. Zero balanced for changes in E_B and r_p. Cannot be cascaded on common bleeder. Gain varies with change in r_p.

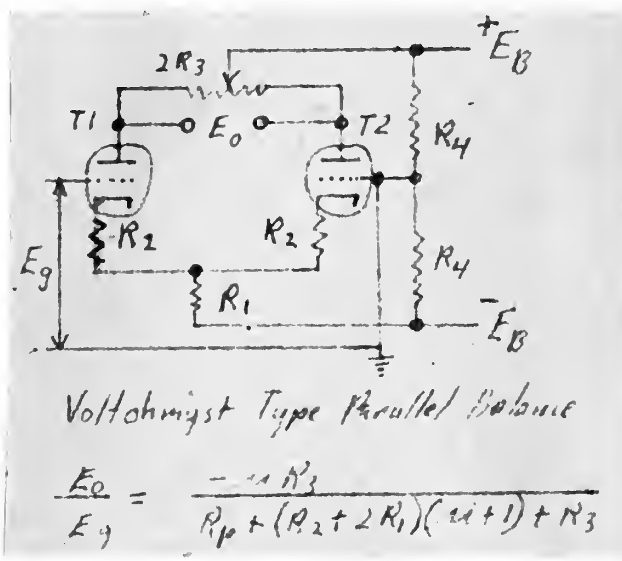


Fig. 11

Low drift. Zero balanced for changes in E_B and r_p. Gain varies with r_p. Not easily overloaded as a meter amplifier.

casr le

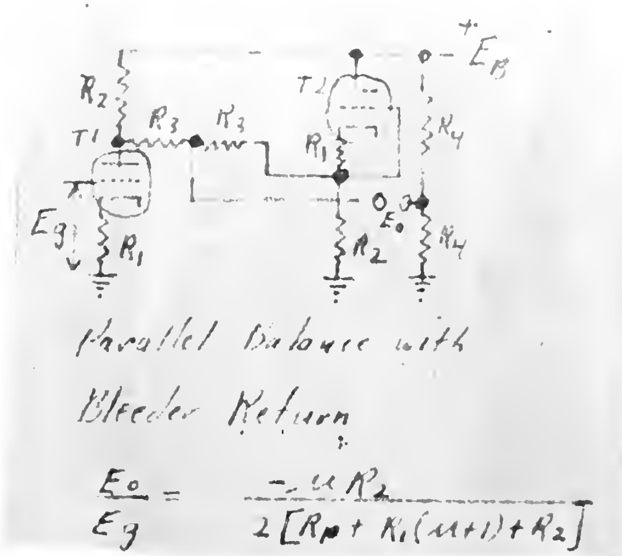


Fig. 12

Drift and balance stability poorer than simple parallel balance type. Can be cascaded on common bleeder. Gain varies with r_p.

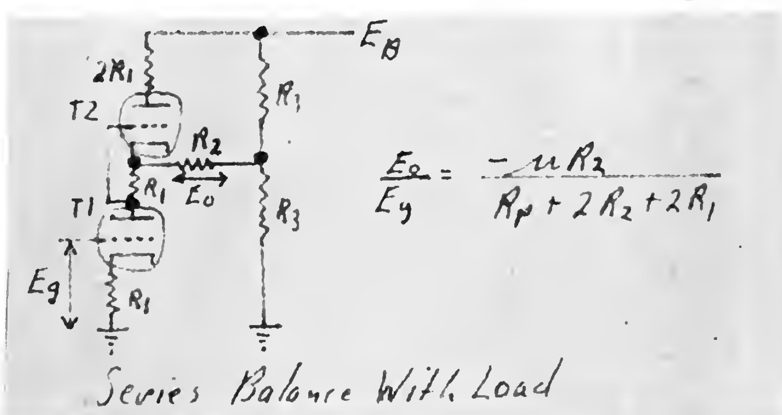
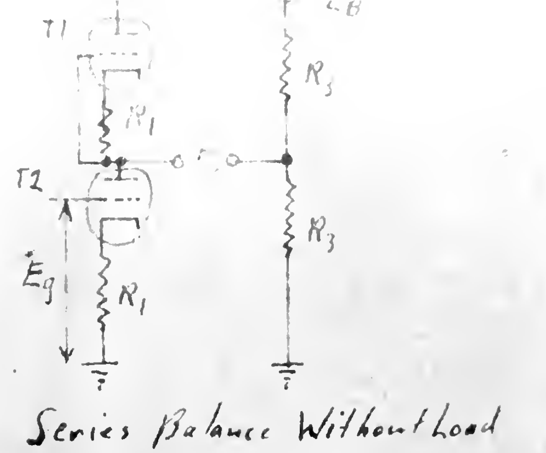


Fig. 13

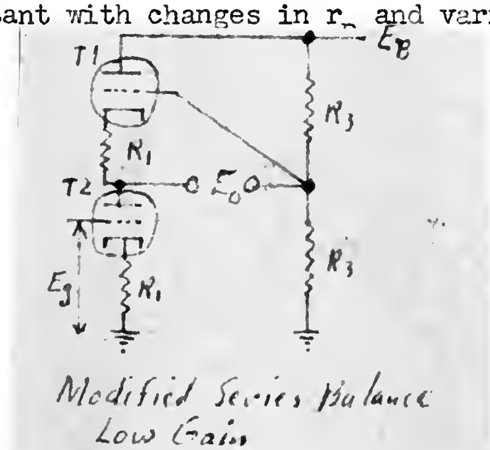
Very low drift. Zero balanced for changes in E_B and r_p. Can be cascaded on common bleeder. Gain varies with r_p.



$$\frac{E_o}{E_g} = \frac{-\mu}{2}$$

Fig. 14

Drift and balance same as for fig. 13. Can be cascaded on common bleeder. Gain constant with changes in r_p and varies with u only.



$$\frac{E_o}{E_g} = \frac{-\mu}{\mu + 2}$$

Fig. 15

Low drift. Zero balanced for change in E_B and r_p. Gain less than unity but constant as r_p varies. Useful as a driver stage.

The above figures 10 through 15 are referred to as bridge balanced in distinction to compensated amplifiers by reason of the fact that all variables in the amplifying arm including tube characteristics and supply voltages are balanced against similar variables in the second arm. The advantage of this type of balance is that regulation of plate and filament supplies is unnecessary except for extreme precision.

The conventional circuit of fig. 10 is the equivalent to returning the output voltage to a bleeder whose current proportioning changes in the same manner as that of the amplifying tube. Fig. 11 shows a circuit essentially the same as that of Fig. 10 but the feedback introduced by the common cathode resistor serves to stabilize the zero of amplifier. Neither of these two can be cascaded and are therefore useful mainly as meter amplifiers. The circuit of Fig. 10 may be altered as shown in Fig. 12 for cascading. Gain per stage is cut in half and the high frequency response is poor due to the high value of resistors R_3 that feed the output. This circuit has a greater drift than that of Fig. 10. Also, for balance, there is the requirement that four pairs of resistors as well as the tubes remain constant in value. The circuits of Fig.'s 13 and 14 eliminate the disadvantages of that of Fig. 12 by replacing the plate load resistor of T_1 and T_2 and its cathode resistor. Balance for E_B and r_p is more nearly perfect than in the Fig. 10 circuit. Differential contact potential changes affect the output but due to the series arrangement the effect is much less pronounced than in other arrangements. In Fig. 13, if R_2 is low in value, load current flow gives regeneration and somewhat higher gain. However if R_2 r_p then there is little difference if R_2 is removed. Therefore the circuit of Fig. 14 is preferred due to lesser number of components. In these circuits the cathode resistor of V1 is made variable for balance adjustment (± 15% from the value of the other cathode resistor). Circuit Fig. 15 is a low gain current amplifier for a driver stage. Drift for this circuit is slightly greater than that for circuits Fig.'s 13 and 14 but very low.

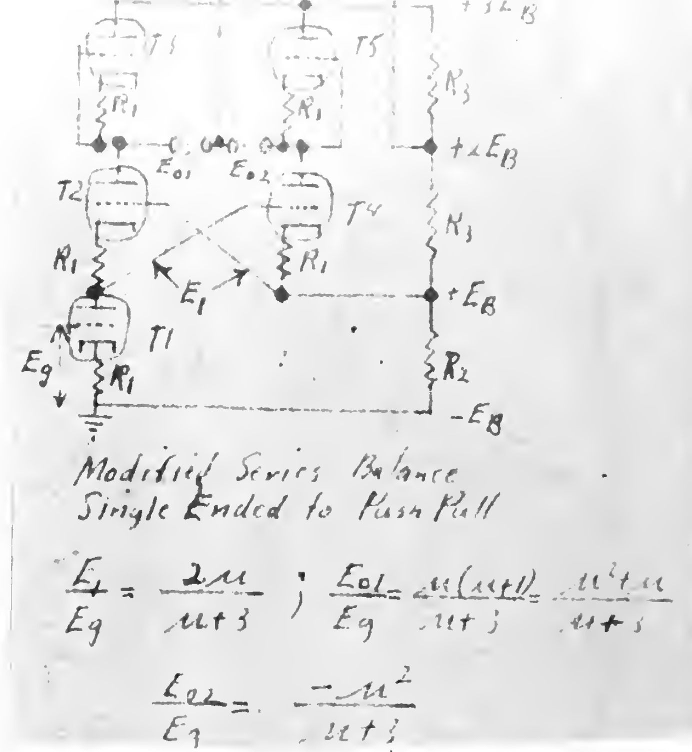


Fig. 16

The circuit of Fig. 16 is a modified form of Fig. 14 to accomplish phase inversion.

The following type circuit is referred to as the cathode follower coupled DC amplifier. The derivation of the mathematical expressions giving the circuit design relationships is as follows:

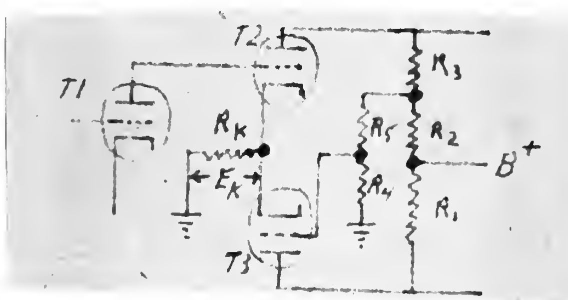


Fig. 17

Notation:	$(E_B - I_{bo2})$	quiescent potential at T_3 grid.
	g_2	transconductance of T_2 at operating point.
	E_B	plate supply voltage.
	i_{p2}	instantaneous signal component of T_2 plate current
	e_{g2}	instantaneous signal component of T_2 grid voltage.

$R_2 + R_3 = R_1$

$$R_k = \frac{E_{bo1} + E_{cc}}{2I_{bo2}}$$

E_{cc} bias voltage of T_2 or T_3.

E_{bo1} quiescent plate voltage of T_1 .

I_{bo2} quiescent plate current of T_2 or T_3 .

$-e_{g2}R_2g_2$ instantaneous signal component of T_3 grid voltage

$$(E_B - I_{bo2}R_2)\frac{R_4}{R_4 + R_5} = E_{bo1} \quad \text{or} \quad \frac{R_4}{R_4+R_5} = \frac{E_{bo1}}{E_B - I_{bo2}R_2} \qquad \underline{1}$$

$$\text{and} \quad \frac{R_4}{R_4+R_5}(e_{g2}\, g_2\, R_2) = e_{g2} \quad \text{or} \quad \frac{R_4}{R_4+R_5} = \frac{1}{R_2\, g_2} \qquad \underline{2}$$

Equating the right hand sides of 1 and 2,

$$R_2 = \frac{E_B}{E_{bb1}\ g_2 + I_{bo2}}$$

The design procedure is;

Choose	Tube type
	E_{bo1}
Compute	R_k
	R_2
Make	$(R_4+R_5) \gg R_2$

The above comprise the basic phase inverting DC amplifier. Figure 18 shows a DC phase inverter amplifier of the noise compensated type.

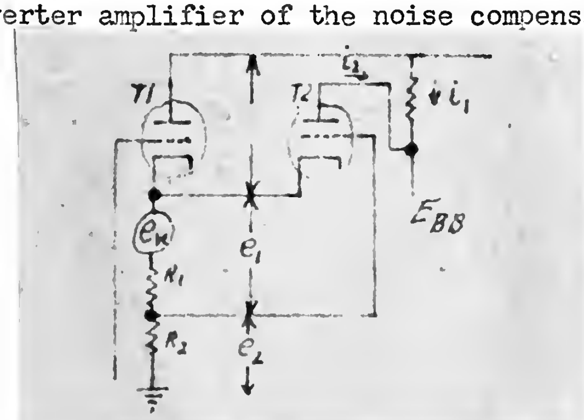


Fig. 18

$E_1 = e_n - (i_1 + i_2)R_2$

$E_2 = (i_1 + i_2)R_2$

For signal input of zero

$i_1 = g_1 (E_1 - E_2)$

$i_2 = g_2 E_1$

$$E_1 - E_2 = \frac{E_1 - g_2 E_1 R_2}{1 + g_1 R_2}$$

if $R_2 = \frac{1}{g_2}$ then $E_1 - g_2 E_1 R_2 = 0$ and $E_1 - E_2 = 0$

or the work function variations of E_1 do not affect i_1 and hence the output. Note that this arrangement does not stabilize for changes in r_p. (This circuitry is very similar to the cathode compensated amplifier mentioned above.) Figure 19 illustrates screen coupled form of the cathode follower DC amplifier.

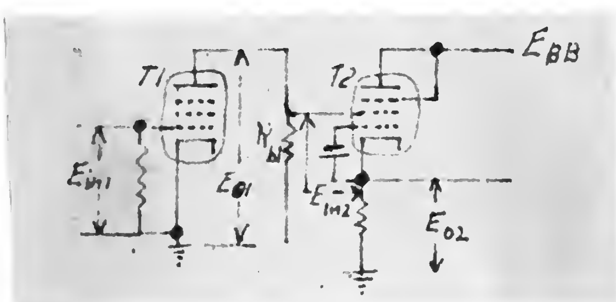


Fig. 19

From elementary equivalent circuit algebraic treatment,

$$E_{o1} = \frac{g_{m1} E_{in1} R_{b1} + \frac{E_{o2} R_{b1}}{R_{sg}}}{1 + \frac{R_{b1}}{R_{sg}}}$$

$$E_{o2} \qquad \frac{g_{m1} R_{b1} g_{2p} E_{in1}}{\frac{1}{R_k} + g_{2p}}$$

Where it is assumed that

$$r_{sg} \gg R_b \quad \text{and} \quad r_{sg} \gg R_k \quad \text{and} \quad \frac{1}{r_{sg}} \ll g_{2p}$$

g_{m1}	control grid to plate transconductance-T_1
R_b	Plate load resistance
g_{2p}	screen grid to plate transconductance-T_2
r_{sg}	screen grid resistance-T_2

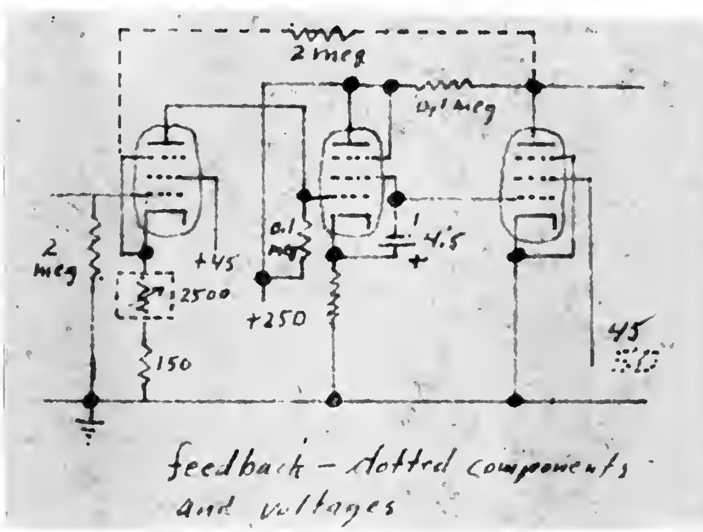


Fig. 20

Figure 20 illustrates a practical amplifier using cathode noise compensation. It has a gain of 76 db. at 15 volts output with a 12 kilocycle bandpass in the absence of feedback. When the feedback loop is inserted it has 60 db. gain at 15 volts out with a 20 kilocycle bandpass.

There exist several types and arrangements of the current coupled or differential amplifier. The "direct connected" type of figure 21 is one of these.

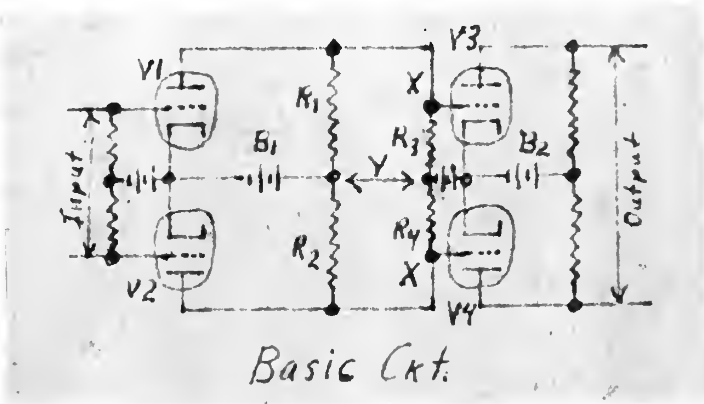


Fig. 21

Note the lack of connection at "Y". This arrangement serves to eliminate the B-supply relation between stages and the high bias to the cathodes of V_3 and V_4. The signal to the second stage is push-pull due to the differential plate current of the first stage, hence the term "current coupled". Further there is no amplification of B-supply variations. Figure 22 illustrates a typical three stage amplifier of this type. The typical differential amplifier is

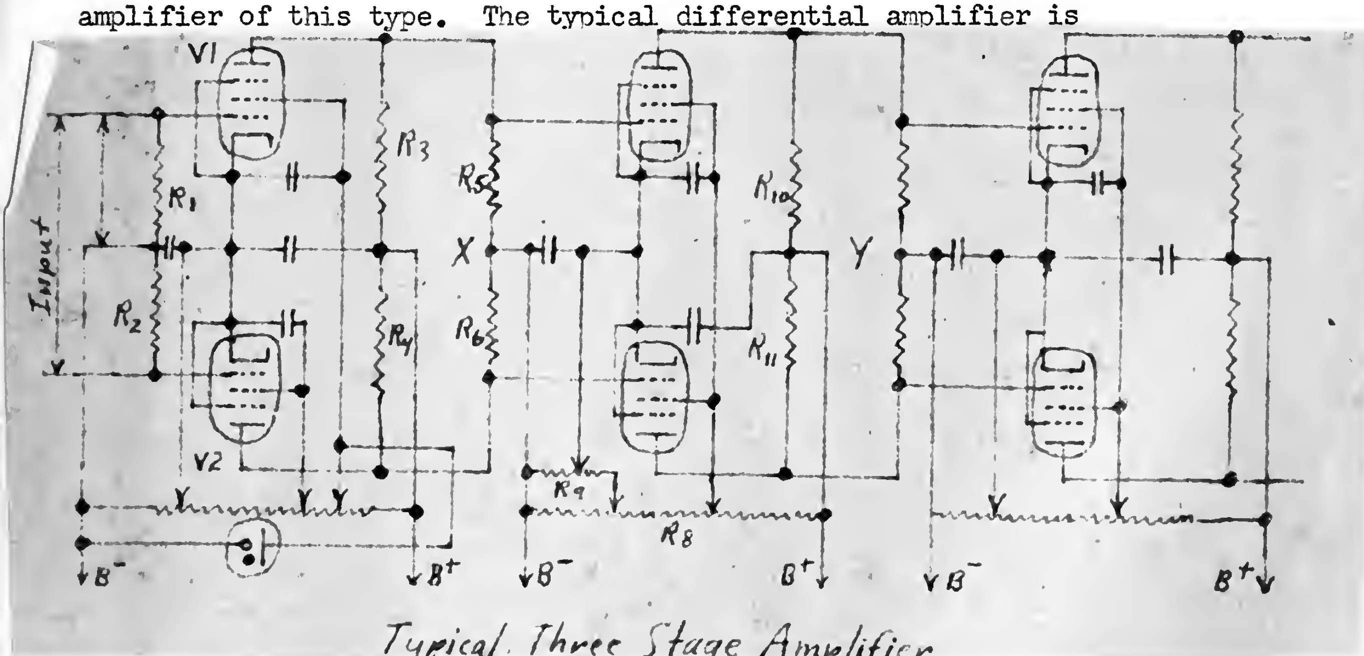


Fig. 22

illustrated in the circuit of figure 23 wherein a common B-supply is used to

feed all stages. In this type of amplifier the operation may

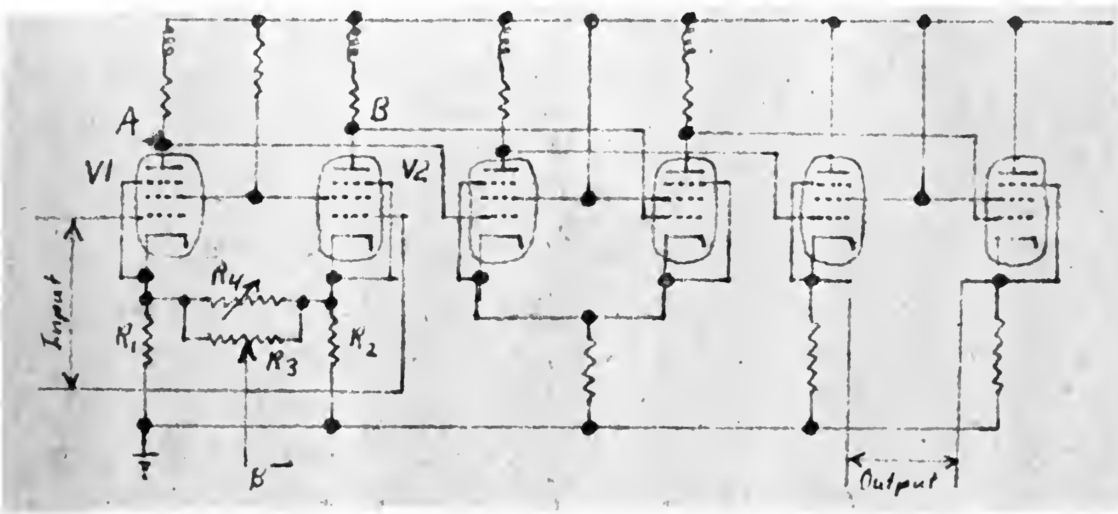


Fig. 23

considered as due to a loop current setup in the loop $V_1R_1R_2V_2$ as a result of the input signal. Note that by the input connected as shown the input DC level may be discriminated against. The design considerations are choice of large values for R_1 and R_2 for degeneration of push-push inputs. The B^- supply is fed to R_1 and R_2 to prevent heavy biasing of V_1 and V_2 such that g_m is reduced. R_4 affects the "loop" current magnitude and hence the gain. The inductors are shunt peaking coils. This same type circuit may be operated in a cathode follower connection as illustrated in the "cross-coupled" amplifier of Figure 24. This circuit has the cathode degeneration advantage of non-criticality of balance adjustment.

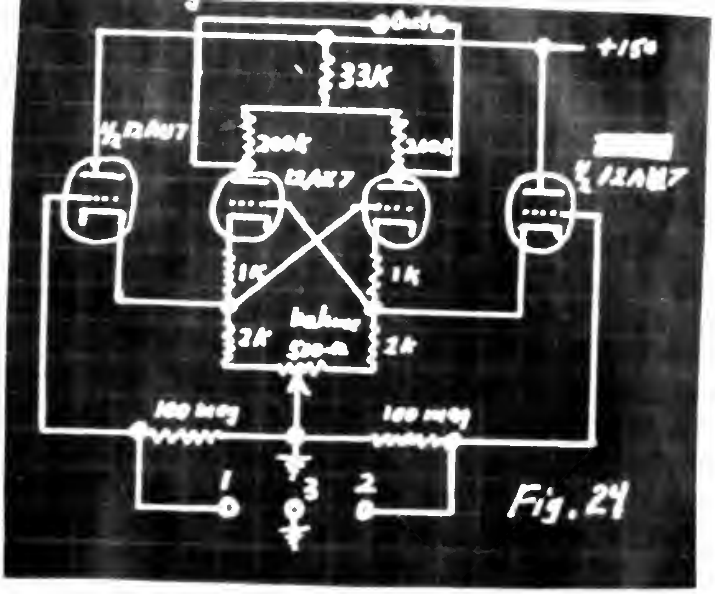


Fig. 24

The basic differential amplifier as it is used today in computer circuitry is drawn in Figure 25. Figure 26 illustrates two arrangements of this amplifier. Figure 27 illustrates several functions which this amplifier can perform. The equations in the illustrations explain the limitations clearly.

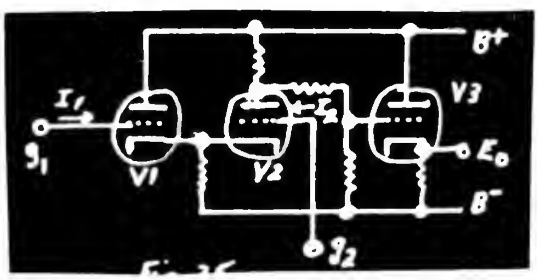


Fig. 25

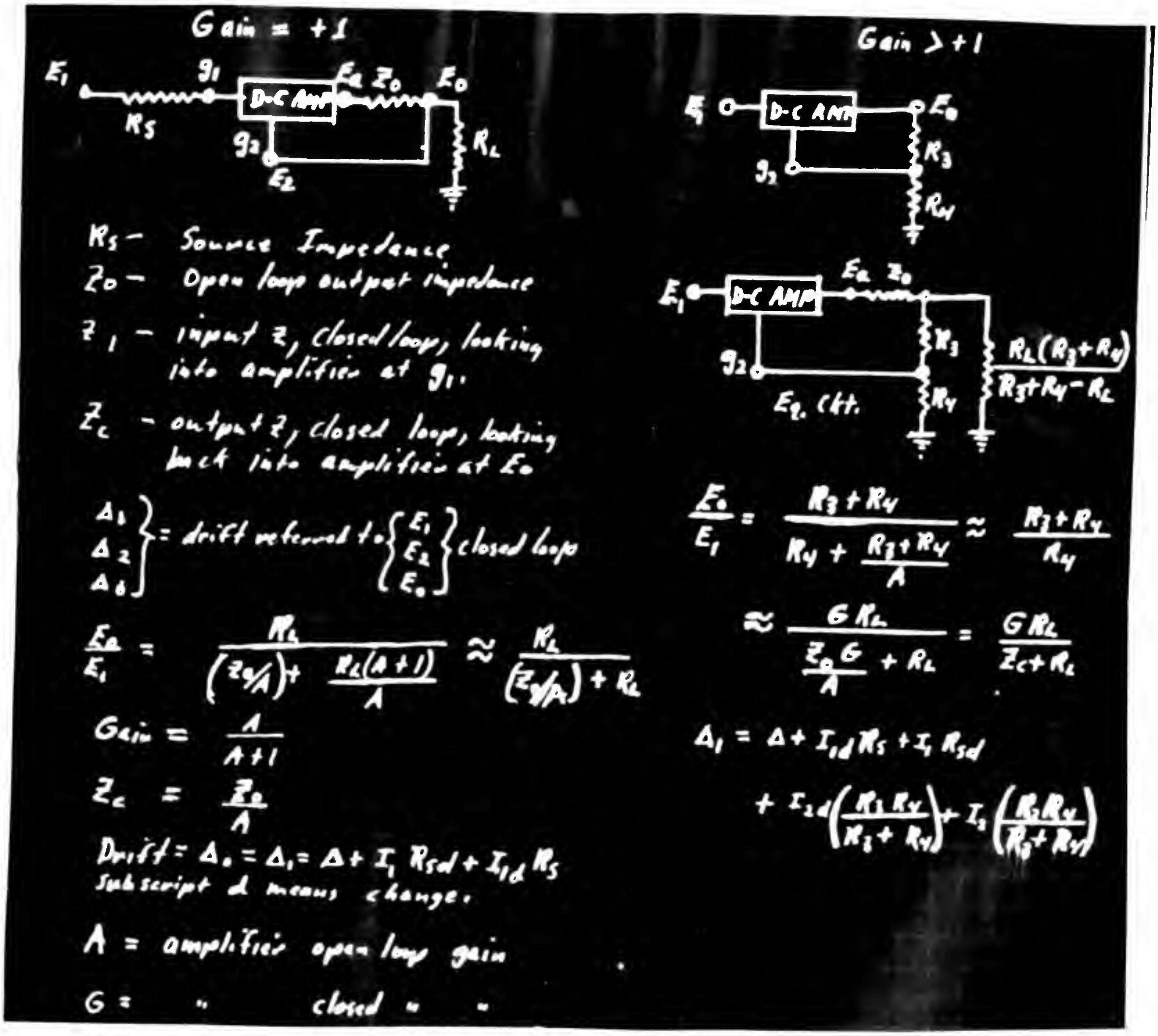
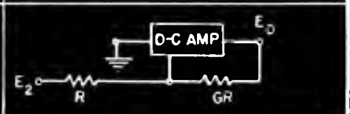
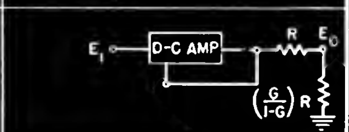
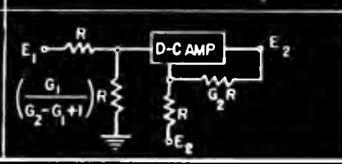

Gain = +1
E_1
g_1
E_a
Z_0
E_0
D-C AMP
R_S
g_2
E_2
R_L
R_S - Source Impedance
Z_0 - Open loop output impedance
Z_1 - input Z, closed loop, looking into amplifier at g_1.
Z_c - output Z, closed loop, looking back into amplifier at E_0
$\left.\begin{matrix}\Delta_1\\ \Delta_2\\ \Delta_0\end{matrix}\right\}$ = drift referred to $\left\{\begin{matrix}E_1\\ E_2\\ E_0\end{matrix}\right\}$ closed loop
$\frac{E_0}{E_1} = \frac{R_L}{(Z_0/A) + \frac{R_L(A+1)}{A}} \approx \frac{R_L}{(Z_0/A) + R_L}$
Gain = $\frac{A}{A+1}$
$Z_c = \frac{Z_0}{A}$
Drift = $\Delta_0 = \Delta_1 = \Delta + I_1 R_{sd} + I_{1d} R_S$
Subscript d means change.
A = amplifier open loop gain
G = " closed " "
Gain > +1
E_1
D-C AMP
E_0
R_3
g_2
R_4
E_1
D-C AMP
E_a
Z_0
R_3
g_2
R_4
Eq. Ckt.
$\frac{R_L(R_3+R_4)}{R_3+R_4-R_L}$
$\frac{E_0}{E_1} = \frac{R_3+R_4}{R_4 + \frac{R_3+R_4}{A}} \approx \frac{R_3+R_4}{R_4}$
$\approx \frac{G R_L}{\frac{Z_0 G}{A} + R_L} = \frac{G R_L}{Z_c + R_L}$
$\Delta_1 = \Delta + I_{1d} R_S + I_1 R_{sd}$
$+ I_{2d}\left(\frac{R_3 R_4}{R_3 + R_4}\right) + I_2\left(\frac{R_3 R_4}{R_3 + R_4}\right)$

Fig. 26

MATHEMATICAL FUNCTION AND CIRCUIT EQUATION	CIRCUIT
(1) MULTIPLICATION BY A NEGATIVE CONSTANT $E_0 = -GE_2$	
(2) MULTIPLICATION BY A POSITIVE CONSTANT LESS THAN ONE $E_0 = GE_1$	
(3) MULTIPLICATION BY A POSITIVE CONSTANT LESS THAN ONE $E_0 = GE_1$	
(4) MULTIPLICATION BY ONE $E_0 = E_1$	
(5) MULTIPLICATION BY A POSITIVE CONSTANT GREATER THAN ONE $E_0 = GE_1$	
(6) MULTIPLICATION OF TWO VARIABLES, ONE BY A POSITIVE CONSTANT AND ONE BY A NEGATIVE CONSTANT $E_0 = G_1 E_1 - G_2 E_2$	

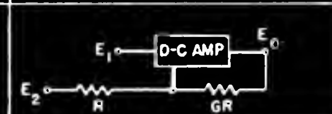
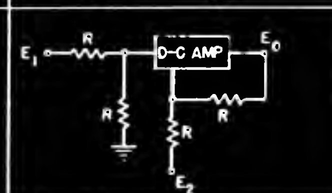
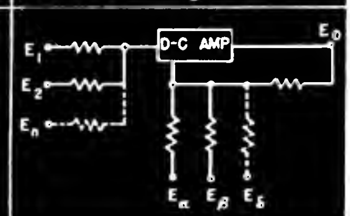
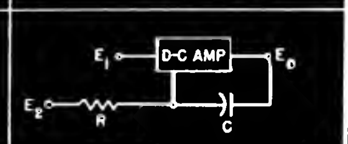
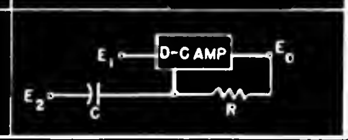
MATHEMATICAL FUNCTION AND CIRCUIT EQUATION	CIRCUIT
(7) SPECIAL CASE OF EXAMPLE 6 $E_0 = (G+1)E_1 - GE_2$	
(8) SIMPLE SUBTRACTION $E_0 = E_1 - E_2$	
(9) MULTIPLICATION OF n VARIABLES, EACH BY A POSITIVE OR NEGATIVE CONSTANT $E_0 = (G_1 E_1 + G_2 E_2 + \ldots G_n E_n) - (G_\alpha E_\alpha + G_\beta E_\beta + \ldots G_\delta E_\delta)$	
(10) INTEGRATION $E_0 = E_1 + \frac{1}{RC}\int_{t=0}^{t} [E_1(t) - E_2(t)]\,dt$	
(11) DIFFERENTIATION $E_0 = E_1 + RC\frac{d}{dt}[E_1(t) - E_2(t)]$	

Fig. 27

When large input signals are available (greater than 50 millivolt) the gas tube coupled amplifier is very well suited to use. Figures 28 through 30 illustrate the basic circuit and the more refined version. The equations in the illustration depict the circuit performance. In figures 28 R_g is a compromise choice since we wish to have R_g r_n and R_n, the DC resistance of the gas tube. The result here is a substantial coupling loss. The circuit of figure 29 is designed to eliminate this coupling loss. The advantage lies in the fact that most of the dcvoltage drop in the grid circuit of T_2 is across the VR tube while most of the signal drop is across R_g. (Note: the impedance of gas tubes increases with frequency above 100 cps. Use of by pass capacitors across the VR tube is recommended.)

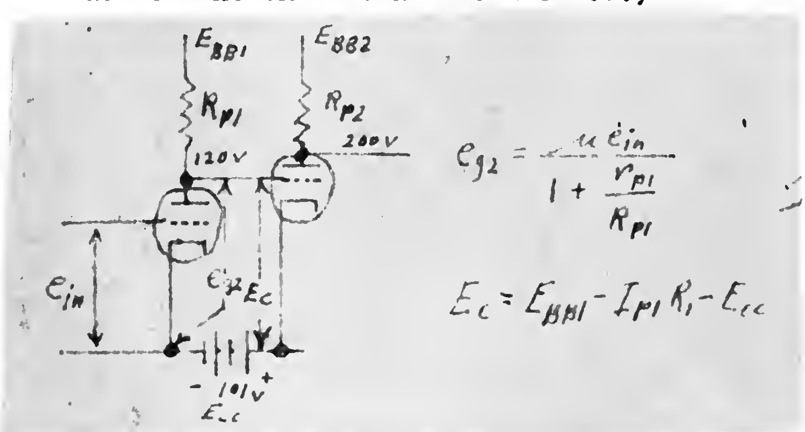


Fig. 28

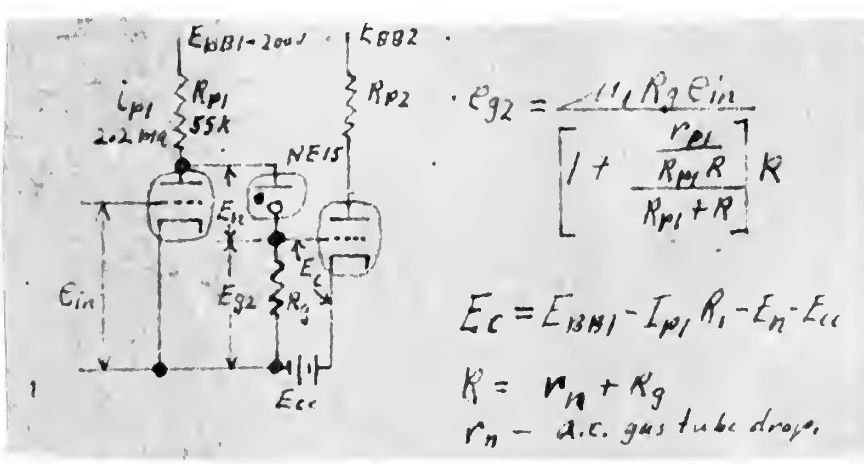


Fig. 29

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